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Efficiently modulate ultrawideband signals

Product Technology
Conductive silicones supply flexible shielding

Clock Translators Upgrade Data Links

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Communications Issue

MILLIMETER WAVE
MIXER ASSEMBLIES

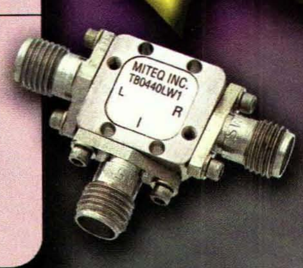
From Stock

MIXERS

Model Number	Frequency (GHz)			LO Power (dBm)	Conversion Loss (dB Typ.)	LO-RF Isolation (dB, Typ.)
	RF	LO	IF			
TB0440LW1	4–40	4–42	.5–20	10–15	10	20
DB0440LW1	4–40	4–40	DC–2	10–15	9	25
SBE0440LW1	4–40	2–20**	DC–1.5	10–15	10	20
IR2640L17*	26–40	26–40	Note 1	15	10	15
M2640W1	26–40	26–40	DC–12	10–12	10	20
TB2640LW1	26–40	26–40	.5–20	10–15	10	20

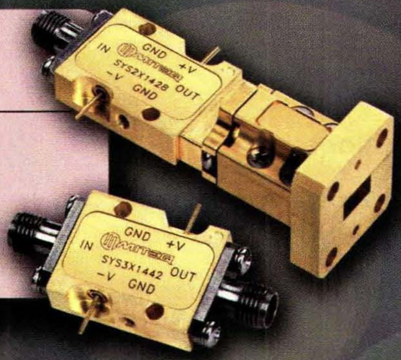
* Image Rejection typically 15 dB. ** Sub Harmonic

Note 1: IF Option A: 20–40 MHz, B: 40–80 MHz, C: 100–200 MHz, Q: DC–1000 MHz

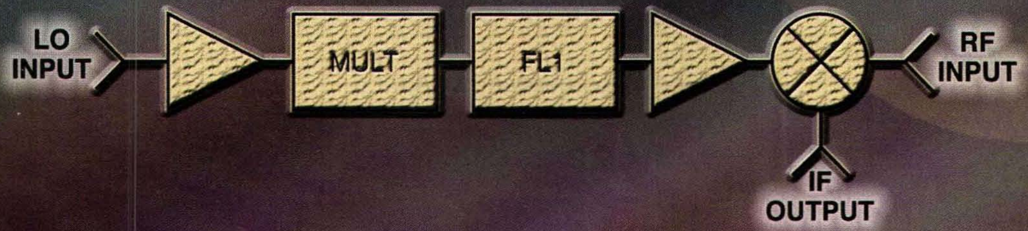


MULTIPLIERS

Model Number	Frequency (GHz)		Input Power (dBm)	Output Power (dBm, Typ.)	Fundamental Leakage (dBc, Typ.)
	Input	Output			
SYS2X1428	14	28	+12	+12	-50
SYS2X1734	16–17.5	32–35	+12	+12	-50
SYS3X1442	14	42	+12	+12	-50
SYS4X1146	11	46	+12	+15	-60
SYS2X2040	10–20	20–40	+12	+15	-15
TD0040LA2	2–20	4–40	+10	-5	-20



MIXER/MULTIPLIER ASSEMBLIES



Model Number	Frequency (GHz)			LO Power (dBm)	Conversion Loss (dB, Typ.)	Input IP^3 (dBm, Typ.)	Fundamental LO-RF Isolation (dB, Typ.)
	RF	LO	IF				
SYSMM2X2335	23.67–35.33	11.385–17.665	.04–.230	13–15	12	+15	50
SYSMM3X2640	26.5–40	8.8–13.3	DC–.5	10	10	+15	40

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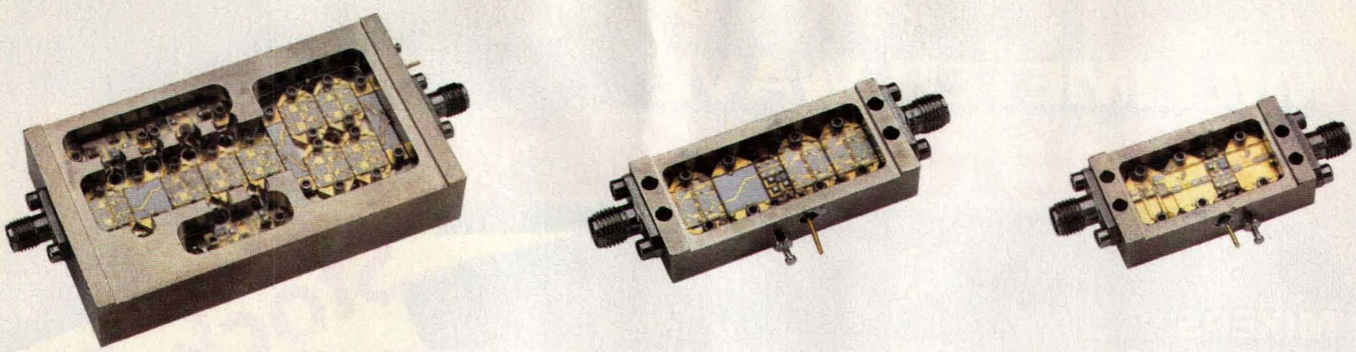

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Ultra broadband

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA018-203	0.5–18.0	20	5.0	2.5	7	17	2.0:1	250
JCA018-204	0.5–18.0	25	4.0	2.5	10	20	2.0:1	300
JCA218-506	2.0–18.0	35	5.0	2.5	15	25	2.0:1	400
JCA218-507	2.0–18.0	35	5.0	2.5	18	28	2.0:1	450
JCA218-407	2.0–18.0	30	5.0	2.5	21	31	2.0:1	500

Multi-octave amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA04-403	0.5–4.0	27	5.0	1.5	17	27	2.0:1	550
JCA08-417	0.5–8.0	32	4.5	1.5	17	27	2.0:1	550
JCA28-305	2.0–8.0	22	5.0	1.0	20	30	2.0:1	550
JCA212-603	2.0–12.0	32	5.0	3.0	14	24	2.0:1	550
JCA618-406	6.0–18.0	20	6.0	2.0	25	35	2.0:1	600
JCA618-507	6.0–18.0	25	6.0	2.0	27	37	2.0:1	800

Medium-power amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-P01	1.35–1.85	35	4.0	1.0	33	41	2.0:1	1000
JCA34-P02	3.1–3.5	40	4.5	1.0	37	45	2.0:1	2200
JCA56-P01	5.9–6.4	30	5.0	1.0	34	42	2.0:1	1200
JCA812-P03	8.0–12.0	40	5.0	1.5	33	40	2.0:1	1700
JCA1218-P02	12.0–18.0	22	4.0	2.0	25	35	2.0:1	700

Low-noise octaveband LNAs

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-3001	1.0–2.0	40	0.8	1.0	10	20	2.0:1	200
JCA24-3001	2.0–4.0	32	1.2	1.0	10	20	2.0:1	200
JCA48-3001	4.0–8.0	40	1.3	1.0	10	20	2.0:1	200
JCA812-3001	8.0–12.0	32	1.8	1.0	10	20	2.0:1	200
JCA1218-800	12.0–18.0	45	2.0	1.0	10	20	2.0:1	250

Narrowband LNAs

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ	VSWR In/Out max	DC Current mA
JCA12-1000	1.2–1.6	25	0.75	0.5	10	20	2.0:1	80
JCA23-302	2.2–2.3	30	0.8	0.5	10	20	2.0:1	80
JCA34-301	3.7–4.2	30	1.0	0.5	10	20	2.0:1	90
JCA56-401	5.4–5.9	40	1.0	0.5	10	20	2.0:1	120
JCA78-300	7.25–7.75	27	1.2	0.5	13	23	2.0:1	120
JCA910-3000	9.0–9.5	25	1.3	0.5	13	23	1.5:1	150
JCA910-3001	9.5–10.0	25	1.4	0.5	13	23	1.5:1	150
JCA1112-3000	11.7–12.2	27	1.4	0.5	13	23	1.5:1	150
JCA1213-3001	12.2–12.7	25	1.4	0.5	10	20	2.0:1	200
JCA1415-3001	14.4–15.4	35	1.6	1.0	14	24	2.0:1	200
JCA1819-3001	18.1–18.6	25	2.0	0.5	10	20	2.0:1	200
JCA2021-3001	20.2–21.2	25	2.5	0.5	10	20	2.0:1	200

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Some Corrections

▶▶THERE WERE SEVERAL errors in Part 6 of my article series ("Matching Loop Antennas To Short-Range Radios," August, p. 72).

The title of the article is a bit misleading in that Part 5 covered the matching of the tapped-capacitor case, and the meat of the Part 6 article is not the matching. It is proving that the inductively tapped-loop antenna is really a transformer system and analyzing that, from which the matching for this particular case does fall out. We made two suggestions for alternate titles.

A sentence was inserted to explain that an exception to closed-path integration is the case of a partially semi-infinite closed path where some segments are so far away as to have negligible impact. We thought that we should say it plainly since that is, in fact, what is done in analyzing Fig. 14.

Eq. 89 is an approximation. The equal sign should be replaced with an approximation sign.

In the paragraph located above the Differential Drivers subhead on p. 82, the word "which" was incorrectly changed to "with."

Farron L. Dacus
RF Architecture Manager
Microchip Technology, Inc.

Another Correction

▶▶THANK YOU FOR publishing my article, entitled "Base-Station Antennas Offer Dual Polarization," in the May issue of *Microwaves & RF* (p. 99). I would like to make a correction to the caption of Fig. 3, at the top of p. 102. At the end of the caption, please replace ". . . single MLA element" with "subarray." The corrected caption should read, ". . . (b) to form a larger array with approximately 3-dB more gain than a subarray."

Dr. Jamal S. Izadian
President
AntennEM Communication, LLC

Editor's Note: Microwaves & RF regrets the errors.

PLEASE COMMENT

Microwaves & RF welcomes mail from its readers. Letters should be typewritten and must include the writer's name and address. The magazine reserves the right to edit letters appearing in "Feedback." Address letters to:

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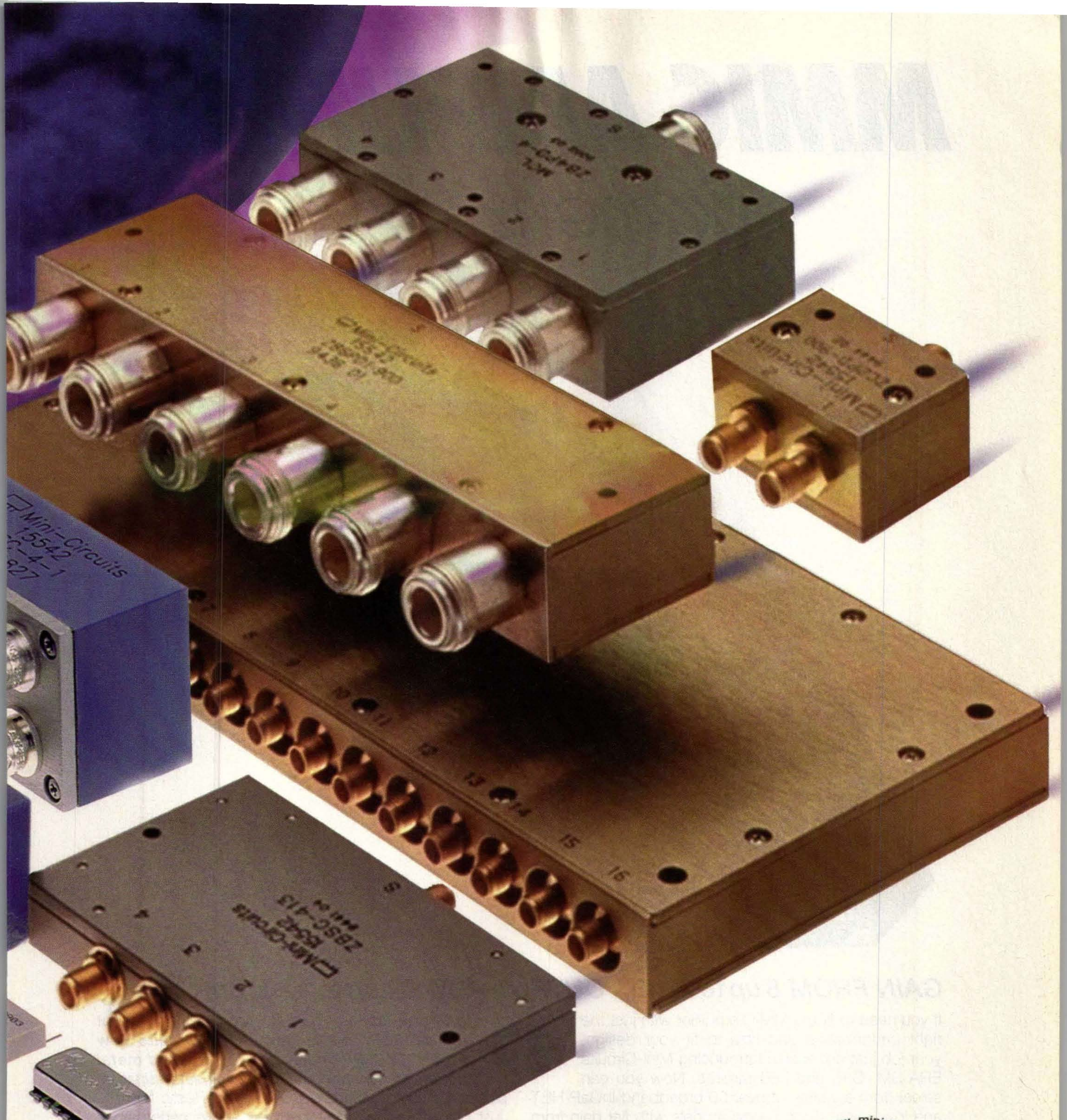

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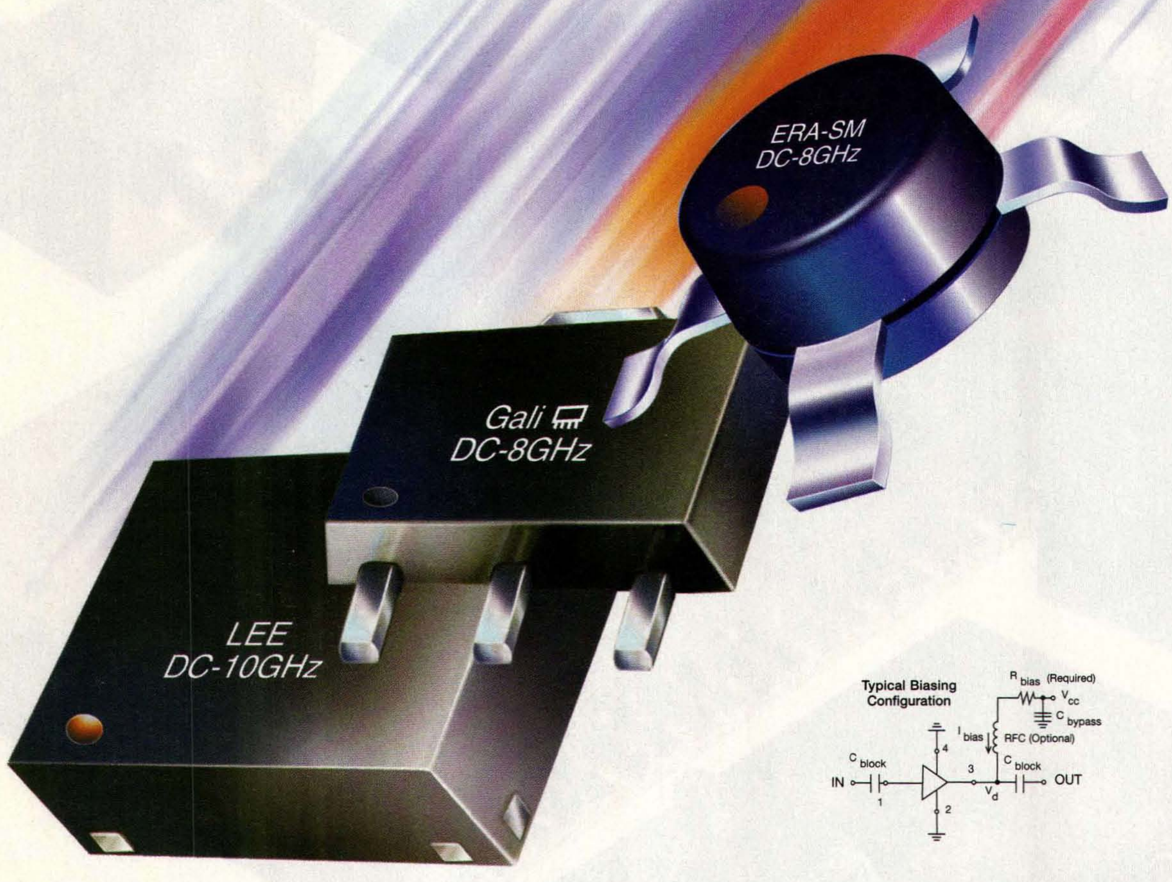

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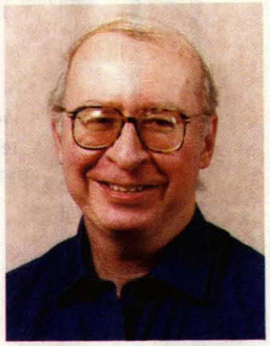
Sampling Military-Electronics Trends



Many high-frequency-electronics manufacturers are once again turning to military customers for future commerce in the hopes of adding positive numbers to next year's revenue projections, thanks to the drastic drop in wireless business, most notably in cellular-related markets over the last several years.

As microwave/RF companies seek to renew business ties in military markets, they are faced with a different set of rules than in the past.

One of the many things that has changed in a decade where RF/microwave companies have departed from their traditional military customers is the name of the customer. Westinghouse, of course, is now Northrop Grumman. The surveillance Rx portion of Watkins-Johnson Co. is now BAE Systems. And once-proud names like Eaton, Emerson Electric, E-Systems, Hazeltine Corp., Martin Marietta, McDonnell Douglas, Sanders, and Sperry have long faded into memory, absorbed into a handful of larger contractors that now dominate the military-electronics landscape.

As microwave/RF companies seek to renew business ties in military markets, they are faced with a different set of rules than in the past. For one thing, military-system designers and maintainers are now forced to compete with their own technologies. For example, the spread-spectrum techniques developed by the military for communications security are now commonplace in cordless consumer telephones and WLANs. For the military to implement their own WLANs, they are faced with the same questions of data security as commercial WLAN designers. And UWB technology, a "carrierless" form of communications that time-codes short pulses to transmit high data rates at low power levels (visit www.uwb.org for more information), is no longer in the exclusive domain of military-system integrators. Military-security experts must plan for a flood of commercial UWB-based chip sets and consumer products as this technology is filtered to commercial markets.

For microwave/RF companies hoping to work with military customers on these problems, there is a place to go for more information—the Military Electronics Show (MES). Scheduled for September 24-25, 2002 in the Baltimore Convention Center, the MES features presentations on WLAN security, UWB technology, power supplies for military use, computers, antennas—in short, on a wide range of technical issues of interest to design engineers involved with military-electronic systems. In addition to the technical conference, a host of leading suppliers of hardware, software, and test equipment will be available in the exhibition hall. For more information, don't miss the MES preview beginning on p. 33, or visit the website at www.mes2002.com.

Jack Browne

Publisher/Editor

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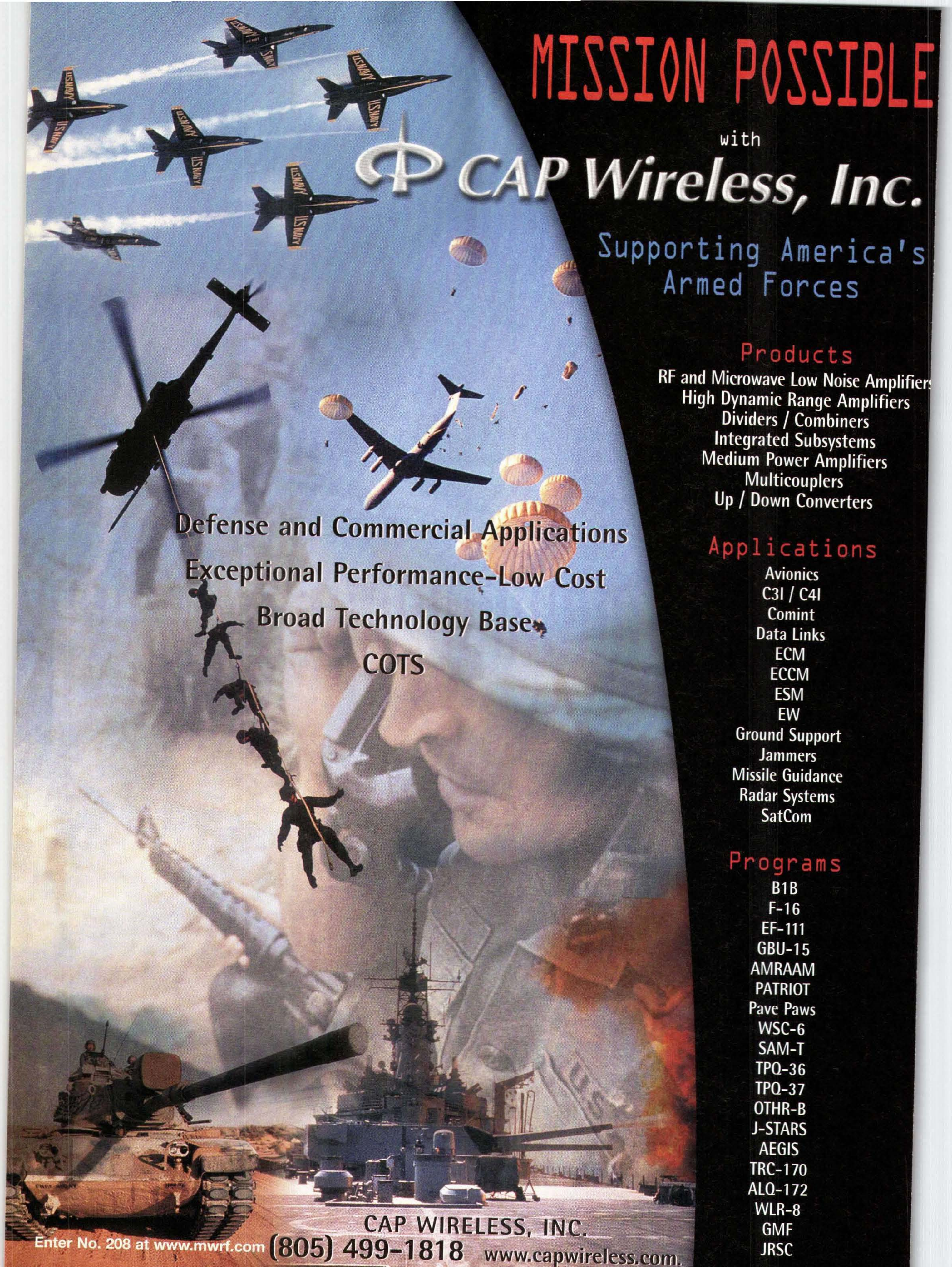
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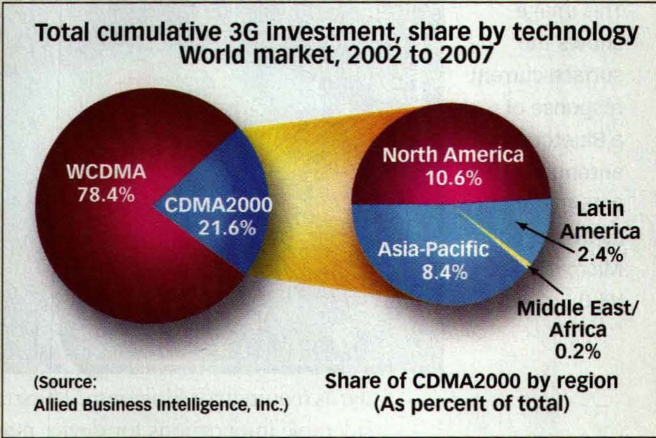
Market Re-evaluation Is Necessary For Wireless Infrastructure

OYSTER BAY, NY—The market for wireless infrastructure has come to a grinding halt. Allied Business Intelligence, Inc. (ABI) estimates that sales of wireless base stations will be down 15 percent this year, with third generation (3G) barely making a dent. Sales of wireless base stations reached approximately $20.3 billion in 2001. This year, sales are expected to decline to $17.4 billion. Due to the downturn, many firms are expected to re-examine their position within the industry.

An ABI report, "Wireless Base Stations: Global Deployments & Revenue for 2G, 2.5G and 3G Systems" by ABI senior analyst Edward Rerisi, examines the trends within the wireless-infrastructure industry and reveals many of the challenges, and opportunities, that lie ahead.

With demand for 2G set to diminish rapidly over the next five years, a greater emphasis will be placed on the rising demand for 3G equipment.



"No one wants to be the first to consolidate or narrow their 3G product portfolio. The market just cannot support as many vendors as we have today," Rerisi stated. Rerisi made a key point regarding the market for 3G equipment based upon code-division-multiple-access (CDMA) technology: Less than a quarter of 3G spending over the next five years consists of cdma2000 equipment **(see figure)**. Geographically, it is almost all in the Americas and Asia Pacific.

Support Is Given For Taiwanese Bluetooth Test Facility

MALAGA, SPAIN—Centro de Tecnologia de las Comunicaciones, S.A. (CETECOM) of Spain and SGS Taiwan Ltd. of Taiwan have announced an agreement to establish a Bluetooth Qualification Test Facility (BQTF) in Taiwan.

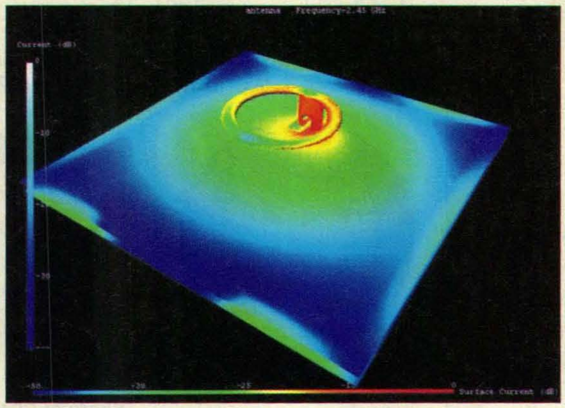
CETECOM Spain will provide technical assistance, consultancy, and training and will also supply test systems for the test facility. The test laboratory will be established using the BITE family of test platforms, which include RF, Protocol, Profile, Engineering, and Blue Unit testers. The plan is to have the facility running and fully operative during the last quarter of this year.

"Taiwan has a growing market of qualified Bluetooth products, being the second country in volume in the Asia Pacific region, after Japan. This new Bluetooth Test Facility will help Taiwanese companies to strengthen their position in the Bluetooth industry and will reinforce the BITE systems as the best testing tools for BQTFs," said Fernando E. Hardasmal, CETECOM's deputy general manager.

"SGS aims to provide to the tremendous Taiwanese IT industry a total solution," commented David Chu, SGS Greater China E&E director. "In the past, most of Taiwanese manufacturers submitted their Bluetooth applications to European and US BQTFs, but that took longer turnaround time and increased the costs. As soon as SGS Taiwan establishes and obtains BQTF accreditation, the Taiwanese manufacturers will immediately take advantage of faster and more economical solutions and this will help them to reduce their manufacturing cost and product lead time.

Research Highlights Bluetooth RF Issues

SOUTHBOROUGH, MA—Research carried out by Flomerics, Inc. has demonstrated the negative effects that plastic enclosures can have on the RF signal performance of Bluetooth antennae. Using the Micro-Stripes V6.0 electromagnetic (EM) modeling software, the research showed that placing a Bluetooth antennae against a plastic enclosure reduced its signal strength by 37.5 percent and shifted the broadcast signal to 2.159-GH—taking the device's performance outside the standards required for Bluetooth. The study has considerable implications for device manufacturers and demonstrates the potential hazards they could face if designers neglect to properly simulate and test their devices' RF outputs 'in-situ.'



This image shows the surface-current response of a Bluetooth antennae at 2.45-GHz input as displayed in Micro-stripes V6.0.

David Johns, vice president for Electromagnetic Engineering at Flomerics, stated, "The research Flomerics has conducted is of consequence to every company that is developing a Bluetooth-enabled device. The magnitude of the change in signal characteristics brought about by simply installing an antenna in a more realistic setting was huge. Manufacturers cannot expect devices to operate optimally simply because a tried-and-tested antenna is being used. Every application will react differently, meaning that simulation and testing of each antenna is required in-situ. If engineers are not aware of these issues before designs are initiated, significant time and money will be wasted building and testing prototypes before an acceptable RF solution is achieved."

Real-world Bluetooth devices often use internal antennae surrounded by plastic enclosures, therefore Flomerics' work centered on a tab-mounted, embedded Bluetooth laptop antenna supplied by a leading manufacturer. To create a realistic setting, a sheet of 3-mm engineering thermoplastic material with the characteristics of a leading plastic was introduced that just touched the top of the antenna. Using Micro-Stripes V6.0 **(see figure)**, a full microwave, RF, and EM simulation showed that the proximity of the thermoplastic shifted the broadcast frequency to outside the Bluetooth range and increased the reflection into the input port by 37.5 percent—severe enough to put the unit's performance outside the limits required for Bluetooth.

In free-space, the Micro-Stripes simulated return-loss for the antenna showed a match of 22 dB centered around 2.428 GHz—well within the Bluetooth range. When the same antenna was simulated with the resin touching the surface, the match dropped by 6 dB or 37.5 percent to 16 dB and was centered around 2.158 GHz—significantly outside the Bluetooth broadcast range. If this had occurred in a production device, the unit would require a significant redesign to bring it up to an actual sensitivity of -70 dBm at 10 m away from the antenna, adding significant costs and lengthening the product's time to market.

Next-Generation SONET/SDH Market To Reach $18B In 2006

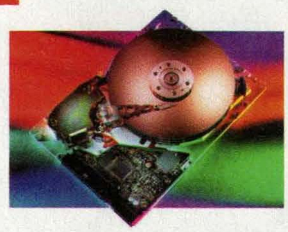
PROVIDENCE, RI—KMI predicts that next-generation Synchronous Optical Network (SONET) and synchronous-digital-hierarchy (SDH) equipment will experience healthy growth from $1 billion in 2001 to nearly $18 billion in annual sales for 2006. Next-generation equipment will largely replace traditional SONET and SDH equipment, opines KMI in the report entitled *Optical Networking: SONET/SDH/Next Generation Equipment*. By 2006, approximately 90 percent of SONET equipment sales will be next-generation equipment.

Several trends have increased the adoption of next-generation equipment over traditional SONET/SDH systems over the past three years, according to KMI. These include the need for better aggregation equipment at the edge of the network, the migration of optical solutions to access networks, cost-competitiveness of next-generation equipment, the move to adopt more efficient technologies, and usage of existing deployments.

KMI Research forecasts that traditional SONET and SDH markets will decline from a high of $17.4 billion in 2000 to $4.6 billion in 2006. The remaining market for SONET and SDH transport equipment will be sales to fill excess capacity on legacy networks.

In addition to the transport market, next-generation equipment will make inroads into the digital cross-connect (DXC) market, the report says.

The products.



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Europe's Incumbent Telcos Need To Define Strategic Options

CAMBRIDGE, ENGLAND—Western Europe's incumbent telecoms operators must look beyond their collective EUR240 billion (approximately $233 billion US) long-term debt and define their strategic options now, states a report from Analysys, an adviser on telecoms and new media.

According to the report, *The Future of Telecoms Incumbents: the impact of competition, regulation, and customer demand*, 14 incumbents had long-term debt at the end of 2001 ranging from around EUR1 billion (approximately $966 million US) to EUR67 billion (around $65 billion US), and half of these operators had gearing of more than 95 percent.

"Management of long-term debt is clearly the principal driver of incumbents' decision making today, as they attempt to restore shareholder value," stated Tamsin Pert, the author of the report. "But they also need to consider how they can create value and maintain competitive advantage in the longer term. In this respect, scenario planning will be key to assessing the potential impact of the decisions they take."

Despite the difficulties faced by incumbent and alternative operators alike, the European telecoms market is still growing (by 9.5 percent in 2001), but it is characterized by unpredictable demand, weak competition, and confused regulation. The success or otherwise of General Packet Radio Service (GPRS) and third-generation (3G) networks will be a crucial factor for most incumbents, many of which are already under financial constraints and have capped capital expenditure for the next few years.

"If GPRS or 3G networks and services do not perform as customers expect, or if costs are higher than planned, there will be huge implications for many incumbents," added Pert.

"If GPRS or 3G networks and services do not perform as customers expect, or if costs are higher than planned, there will be huge implications for many incumbents."

Near-Term Demand For Point-To-Point RF Links Will Stay Low

LONDON, ENGLAND—A report from the Gallium Arsenide and High Speed Circuits service (GaAs) of Strategy Analytics, the global research and consulting company, notes that there will be no increase in near-term demand for point-to-point RF links.

The report predicts that future demand will be driven by renewed network expansions outside of North America and Europe. A number of factors make the less-developed world a more fertile field for RF deployment including high capital cost for wire-line construction, geographic constraints, and limited competition for RF spectrum.

"The difficulties of doing business in the less-developed world are not trivial," noted Stephen Entwistle, director of the Strategy Analytics GaAs Service. "However," Entwistle added, "an effective presence in these alternative markets will be vital to long-term growth."

This report, "The Point-To-Point RF Market: A Strategy Analytics Perspective," discusses the implementation of point-to-point links in mobile wireless, fixed wireless broadband, and metropolitan-area networks, outlining the opportunities and drawbacks within each market segment. It makes recommendations for RF component suppliers and point-to-point RF manufacturers.

Kudos

LAFOX, IL—Richardson Electronics, a global provider of engineered solutions for the RF and wireless market, announced that its North American Center of Excellence, TRL Technologies, Inc., has been awarded QS-9000 registration from Underwriters Laboratories, Inc. (UL).

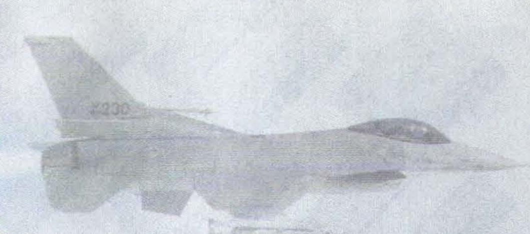
The goal of QS-9000 is to ensure that quality systems are developed to provide continuous improvement, emphasizing defect prevention and the reduction of variation and waste in the supply chain.

WASHINGTON, DC—The Jury for the 2002 Prince of Asturias Award for Technical and Scientific Research has given an award to Dr. Robert Kahn for his efforts in developing the Internet.

The award announcement notes that Kahn is the joint inventor of the TCP/IP protocols and was responsible for setting up the Defense Advanced Research Project Agency (DARPA) Internet program.

The Prince of Asturias Awards recognize and reward "scientific, technological, cultural, social, and humanistic work performed by individuals, groups, or institutions worldwide." The Prince of Asturias Foundation was created in 1980 in the course of a ceremony presided over by the King and Queen of Spain, Don Juan Carlos and Dona Sofia. MRF

Elcom Technologies



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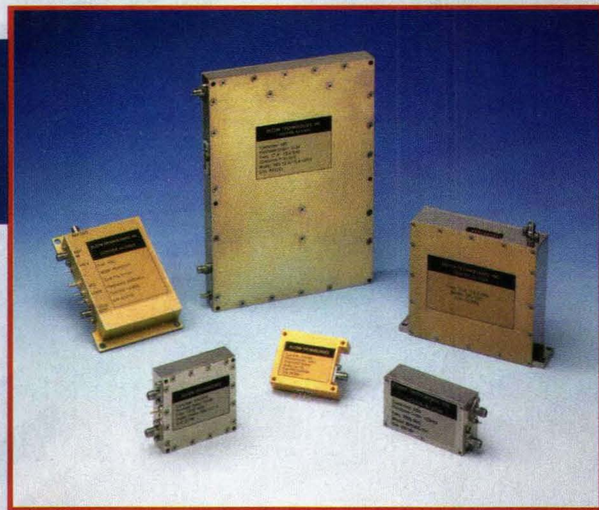

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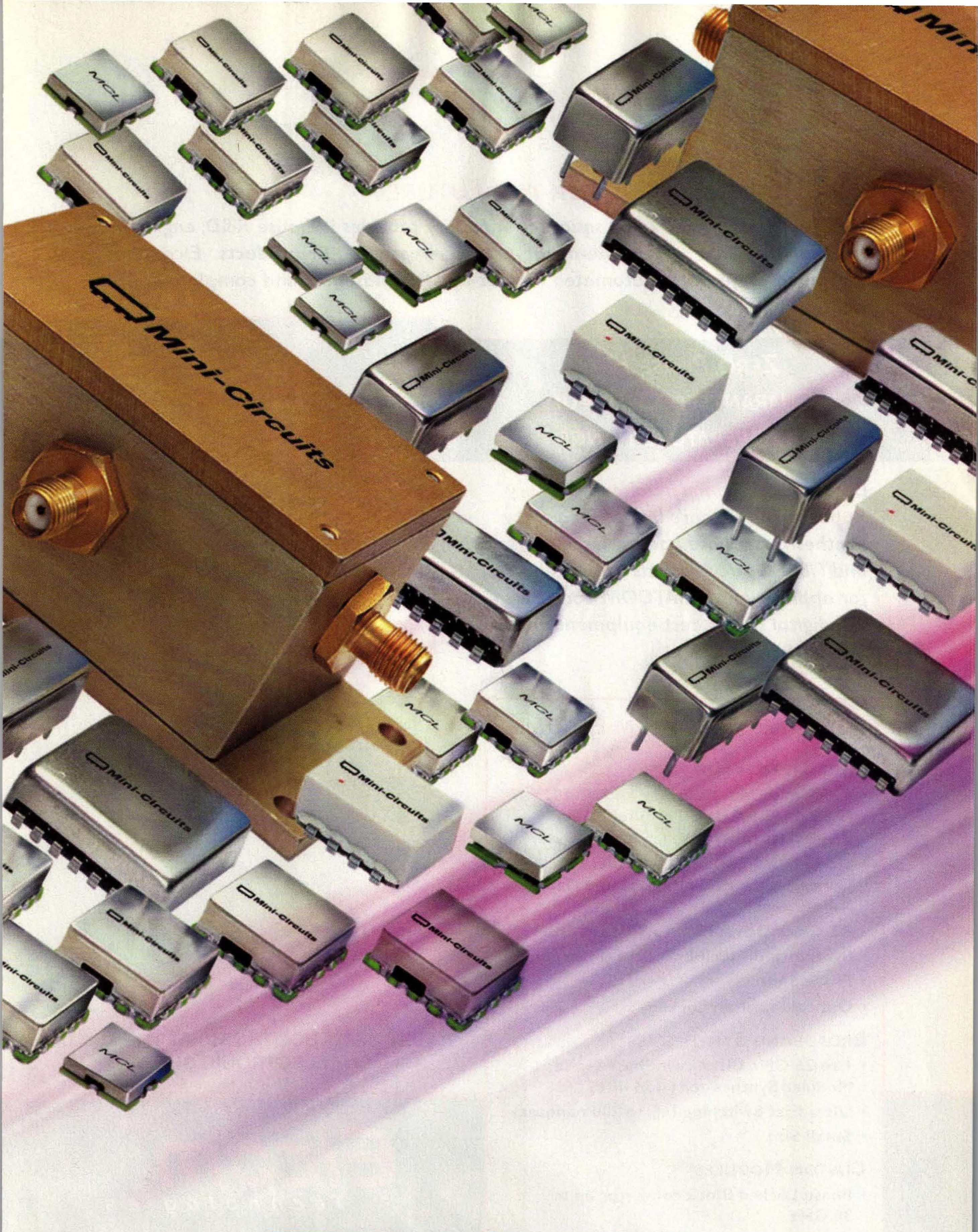
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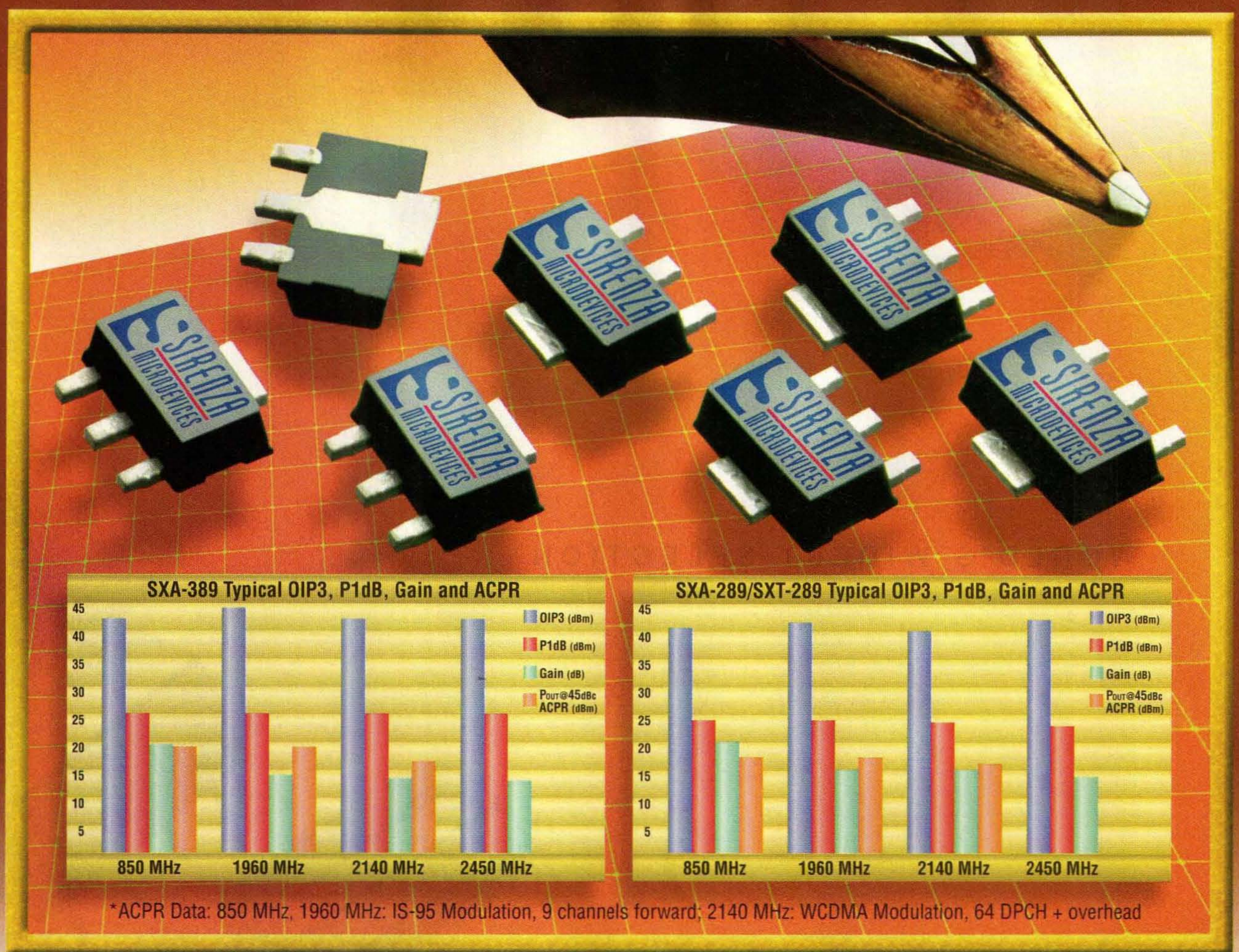

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NEWS

Conference Targets Military Electronics

The second annual Military Electronics Show will offer a wide variety of exhibited products and an extremely educational technical conference for design engineers.

JACK BROWNE
Publisher/Editor

Military applications are once again attracting the interest of high-frequency designers. With the softening of wireless markets, the microwave industry has begun to renew its involvement with military-electronics customers, making this an opportune time for a visit to the industry's only technical conference devoted to true, military-grade high-frequency design. The Military Electronics Show (MES), now in its second year, is scheduled for September 24-25, 2002 at the Baltimore Convention Center (Baltimore, MD). Attendees will learn a variety of skills to help them in working with military customers, including how to make their wireless local-area networks (WLANs) more secure from the threat of data theft, how to evaluate coaxial-cable assemblies for airborne environments, how to use solid-state flask disks for data storage, how to evaluate the latest high-speed interconnection standards for computers and peripheral devices, and even how to create an effective radar simulator using commercial-off-the-shelf (COTS) vector-signal generators.

The technical conference includes presentations on antennas, analog-signal processing/digital-signal processing (ASP/DSP), components, computer-aided engineering (CAE), cables and connectors, electromagnetic interference (EMI)/TEMPEST, packaging and materials, power supplies and converters, receiver (Rx) design, test and measurement, and transmitter (Tx) design. For example, consultant Steve Best, a longtime presenter at the Wireless Symposium & Exhibition, will offer insights into "Antenna Requirements for Ultrawideband (UWB) Systems." Best will examine the types of antennas required for UWB systems, and address some properties of antennas that may be critical to the performance of a UWB system.

Also in the antenna track, Paul Dixon of Emerson & Cuming will educate attendees on the Luneberg Lens. In his talk, entitled, "A Broadband, High-Gain Steerable Luneberg Lens," Dixon will explain this novel technology and show the performance of actual designs. The Luneberg Lens combines the high gain of a dish reflector with the steering ease of a phased array, at considerably less cost.

In an fascinating overview of radar and sonar markets, long-time developer of instantaneous-frequency-measurement (IFM) Rxs and systems, William Sullivan of Wide Band Systems, will review markets for radar and sonar systems and components, from hardware and software perspectives. He will outline the steps that need to be taken by the industry to implement the new tech-

nologies, and identify the obstacles that delay adoption of new technologies.

In a track devoted to cable and connector issues, Thomas Ricard of EZ Form Cable Corporation will talk on "Notch and Comb Filtering Implementations of Coaxial Cable Sections." As Ricard will point out, coaxial-cable implementations of notch and combline filters can be incorporated even in those systems where real estate is at a premium.

In the same cable and connector track, Ted Prema of Times Microwave Systems, in a talk entitled "Coaxial Transmission in Airborne Environments," addresses some of the challenges in specifying coaxial cables for airborne systems. He notes that MIL-C-17 cables with MIL-C-39012 connectors, which are standard for land-based and shipboard military systems, were found to be inadequate for airborne environments. The high failure rate of these cables in this rugged environment resulted in the establishment of specifications MIL-T-81490 and MIL-C-87104 for microwave transmission lines. These specifications address the entire transmission-line assembly, including hermetic sealing, flex life, electrical performance over broad frequency ranges, and mechanical ruggedness.

At the first Wireless Symposium & Exhibition in San Jose, CA in 1993, visitors from the Department of Defense (DoD) gave a presentation on COTS in military systems. To close the loop on this first military/wireless/military cycle, Al Steel of Texas Instruments offers a presentation, "Mitigating the Risks that OEMs Face When Using COTS Products in Military Applications," that once again addresses the use of COTS components in military systems. Steel will outline the qualification procedures necessary to bring military-grade plastic-packaged products to market.

For military designers faced with achieving high frequency stability, Shankar Joshi of Synergy Microwave will offer insights into the company's development of phase-locked-loop (PLL) and fractional-N frequency synthesizers. These low-phase noise designs achieve low spurious and harmonic performance in compact footprints, without sacrificing current consumption. Uri Yaniv of Elcom Technologies will also make a presentation on high-speed frequency synthesizers for military applications, reviewing a line of designs with submicrosecond switching speed and low phase noise over broad bandwidths.

In a session on meeting EMI and TEMPEST requirements, Ryan Maley of Equipto Electronics will offer a presentation entitled "Design Considerations in Building Shielded Enclosures." The presentation will review a variety of considerations in the design of shielded enclosures, including material selection, gasketing technologies,

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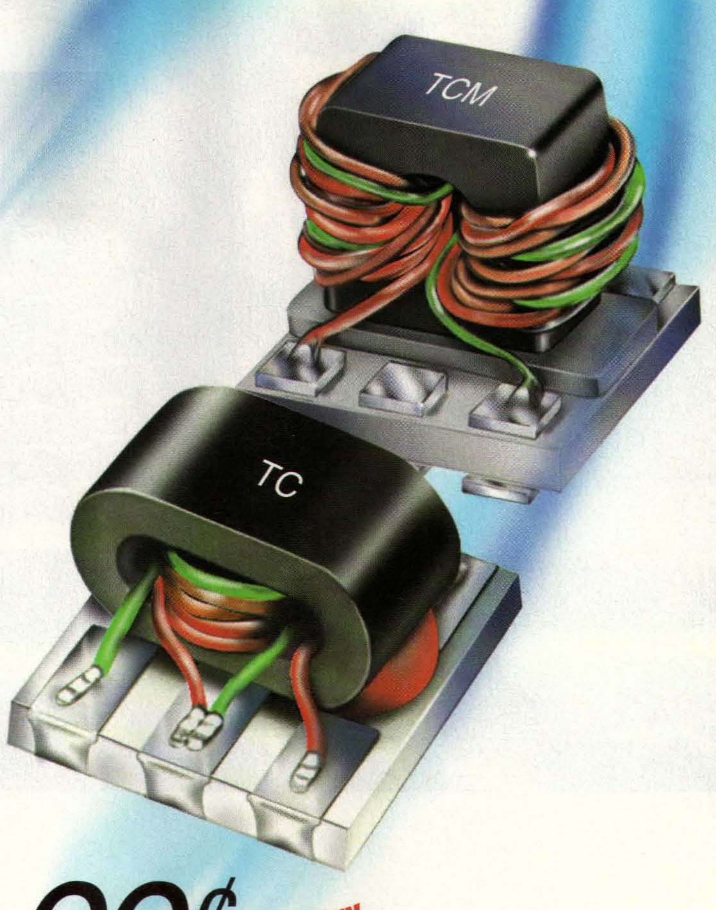
RF TRANSFORMERS



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TCM1-1	1C	1.5-500	5-350	.99
TCML1-11	1G	600-1100	700-1000	1.09
TCML1-19	1G	800-1900	900-1400	1.09
TCM2-1T	2A	3-300	3-300	1.09
TCM3-1T	3A	2-500	5-300	1.09
TTCM4-4	4B	0.5-400	5-100	1.29
TCM4-1W	4A	3-800	10-100	.99
TCM4-6T	4A	1.5-600	3-350	1.19
TCM4-14	4A	200-1400	800-1000	1.09
TCM4-19	4H	10-1900	30-700	1.09
TCM4-25	4H	500-2500	750-1200	1.09
TCM8-1	8A	2-500	10-100	.99
TCM9-1	9A	2-280	5-100	1.19

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Model	Ω Ratio & Config.	Freq. (MHz)	Ins. Loss❖ 1dB (MHz)	Price $ea. (qty. 100)
TC1-1T	1A	0.4-500	1-100	1.19
TC1-1	1C	1.5-500	5-350	1.19
TC1-15	1C	800-1500	800-1500	1.29
TC1.5-1	1.5D	.5-2200	2-1100	1.59
TC2-1T	2A	3-300	3-300	1.29
TC3-1T	3A	5-300	5-300	1.29
TC4-1T	4A	.5-300	1.5-100	1.19
TC4-1W	4A	3-800	10-100	1.19
TC4-14	4A	200-1400	800-1100	1.29
TC8-1	8A	2-500	10-100	1.19
TC9-1	9A	2-200	5-40	1.29
TC16-1T	16A	20-300	50-150	1.59
TC4-11	50/12.5D	2-1100	5-700	1.59
TC9-1-75	75/8D	0.3-475	0.9-370	1.59

Dimensions (LxW): TCM .15" x .16" TC .15" x .15" ❖Referenced to midband loss.

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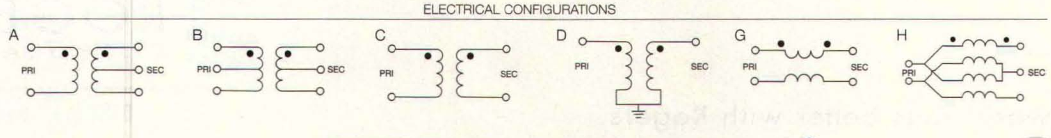

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and closure mechanisms.

In the packaging track, Alan Linder of StratEdge Corporation, in his presentation "High Performance Microwave Packages for Military Applications," will offer an alternative packaging technique to conventional metal-box technology traditionally used in military systems. In addition, Jerry Jordan of Palomar Technologies will offer the presentation "Flip Chip Connections Using Gold Stud Bumps."

Radha Setty of broad-line component supplier Mini-Circuits will review reliable, low-cost components for the military market. In this discussion, he will focus on low-temperature-cofired-ceramic (LTCC) technology, a high-frequency multilayer circuit-board-fabrication technique that supports the use of embedded active and passive devices. Setty will examine the reliability of the technology, as well as qualification testing for individual components.

In a track on test and measurement, Todd Schuler of Sytex, Inc. will evaluate threats and risks associated with the use of IEEE 802.11 WLANs. These networks, which operate in unlicensed frequency bands at 2.4 and 5.0 GHz, are based on various forms of spread-spectrum techniques (originally developed by the military). Schuler will cover the risks faced by users of WLANs, as well as the countermeasures that can be used to protect data transmissions and make WLANs more secure.

Also in the test-and-measurement track, Don McCook of TestMart will examine "Life-Cycle Support of Critical General Purpose Electronic Test Equipment (GPETE)." By providing access to the broad range of available test equipment available through secondary-market test-equipment suppliers through the Internet, exact form, fit, and function replacement instruments can be provided to maintain the readiness of key military support systems.

The test-and-measurement track also features Mike Lauterbach of LeCroy Corporation, who will present "Data Acquisition and Analysis" on new technologies for high-speed data acquisition (DAQ) and analysis. He will offer tips and practical techniques for increasing the throughput of the analysis of complex signals and show how to increase accuracy when digitizing high-speed signals.

Rick Fornes of Besser Associates offers two presentations on receiver design, "Receiver Architectures for Hostile Communication Environments" and "Advanced Receiver Technologies for the 21st Century." In the Tx-design track, Ed Niehenke of Besser Associates will offer an historical sketch of microwave integrated circuits (MICs), monolithic MICs (MMICs), and their application in military systems, such as radar, electronic-warfare (EW) systems, electronic-countermeasure (ECM) systems, and

communications systems. His presentation will include a variety of Tx examples ranging from S-band to W-band.

Niehenke will also perform duties at the Military Electronics Show as invited Keynote Speaker, addressing some of the changes faced by systems houses, as well as companies who are part of the military supply chain. He will give a second technical presentation, "Transmitter Power Combining Techniques."

In the same Tx track, Dr. Clive Winkler of Cubic Communications will discuss "Linearization Methods for Providing High Data Rates Across Existing Radio Frequency Channels." He will describe a feasible method of providing the necessary enabling RF technology so that Txs and Rxs can be linearized for the transmission and reception of more complex symbols at higher data rates. The goal of the presentation is to make Txs and Rxs more compatible with modern Internet Protocol (IP)-addressable devices that support network and file transfers at the media-access-control (MAC) layer level relevant to mobile operations.

Another Tx design session, by Lawrence Burke of M/A-COM, is called "A Telemetry Transmitter Chip Set and Module for Munitions Applications." His talk will review the system architecture of this chip set and module and examine the performance of the individual components.

In a track devoted to computers and peripheral devices, Steve Paavola of Sky Computers will examine the InfiniBand standard and detail how it compared to other computer/peripheral-interconnection standards currently under development. Also in the track on computers and peripheral devices, Roni Kornitz of M-Systems will present his talk "Rugged and Reliable Data Storage: Solid-State Flask Disks." He will explore the use of ruggedized hard disks and solid-state Flash memory disks designed specifically for military and aerospace systems.

In a track on power supplies and converters, Dal Ohm of DriveTech, Inc. will present "Commutation and Current Control Methods for Military Brushless Motors." He will elaborate on the use of current control to improve the performance and reliability of ruggedized military brushless motors. Also in the power-supplies/converters track, Toshio Takashaki of International Rectifier will present "New Digital Hardware Control Method for High Performance AC Servo Motor Drive." He will review the status of modern digital servo drives and introduce a solution for demanding servo drive applications based on his company's high-performance servo drive platform.

Roy Kampmeyer of Power Electronic Systems will offer the presentation "A High Reliability Ground Fault Circuit Interrupter (GFCI) Designed

Ultra-Low Noise AMPLIFIERS

VHF To V-BAND

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMP. (dBm, Min.)	DC POWER @ +15 V (mA, Nom.)
OCTAVE BAND AMPLIFIERS								
JS2-00500100-045-5A	0.5 – 1	35	1	0.45	2:1	2:1	5	250
JS2-00500100-12-5A	0.5 – 1	35	1.2	1	2:1	2:1	5	250
JS2-01000200-045-5A	1 – 2	33	1	0.45	2:1	2:1	5	250
JS2-02000400-045-5A	2 – 4	28	1.2	0.45	2:1	2:1	5	175
JS2-04000800-08-0A	4 – 8	22	1.2	0.8	2:1	2:1	0	150
JS3-04000800-08-5A	4 – 8	30	1	0.8	2:1	2:1	5	175
JS3-04000800-15-5A	4 – 8	30	1	1.5	2:1	2:1	5	175
JS2-08001200-11-5A	8 – 12	15	1	1.1	2:1	2:1	5	150
JS3-08001200-11-5A	8 – 12	25	1	1.1	2:1	2:1	5	175
JS3-08001200-15-5A	8 – 12	25	1	1.5	2:1	2:1	5	175
JS3-12001800-16-5A	12 – 18	23	1	1.6	2:1	2:1	5	175
JS4-12001800-145-5A	12 – 18	30	1	1.45	2:1	2:1	5	200
JS4-12001800-30-5A	12 – 18	30	1	3	2:1	2:1	5	200
JS2-18002600-20-5A	18 – 26	14	2	2	2.5:1	2.5:1	5	100
JS2-18002600-30-5A	18 – 26	14	2	3	2.5:1	2.5:1	5	100
JS3-18002600-20-5A	18 – 26	22	1.8	2	2.5:1	2.5:1	5	175
JS3-18002600-30-5A	18 – 26	22	1.8	3	2.5:1	2.5:1	5	175
JS4-18002600-19-5A	18 – 26	33	1.5	1.9	2:1	2:1	5	200
JS4-18002600-26-5A	18 – 26	33	1.5	2.6	2:1	2:1	5	200
JS2-26004000-35-5A	26 – 40	10	2	3.5	2.5:1	2.5:1	5	100
JS2-26004000-45-5A	26 – 40	10	2	4.5	2.5:1	2.5:1	5	100
JS3-26004000-35-5A	26 – 40	18	2.5	3.5	2.5:1	2.5:1	5	175
JS3-26004000-45-5A	26 – 40	18	2.5	4.5	2.5:1	2.5:1	5	175
JS4-26004000-40-5A	26 – 40	23	2.5	4	2:1	2:1	5	200
JS4-40006000-65-0A	40 – 60	15	3	6.5	2.75:1	2.75:1	0	175
MULTIOCTAVE BAND AMPLIFIERS								
JS2-00500200-07-5A	0.5 – 2	32	1	0.7	2:1	2:1	5	295
JS2-00500200-15-5A	0.5 – 2	32	1	1.5	2:1	2:1	5	295
JS2-01000400-08-5A	1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-01000400-20-5A	1 – 4	27	1	2	2:1	2:1	5	200
JS2-02000600-08-5A	2 – 6	22	1	0.8	2:1	2:1	5	125
JS2-02000600-20-5A	2 – 6	22	1	2	2:1	2:1	5	125
JS2-02000800-08-0A	2 – 8	22	1.25	0.8	2:1	2:1	0	125
JS2-02000800-20-0A	2 – 8	18	1.25	2	2:1	2:1	0	125
JS3-02001800-25-5A	2 – 18	23	1.8	2.5	2.5:1	2.5:1	5	150
JS3-02001800-50-5A	2 – 18	23	1.8	5	2.5:1	2.5:1	5	150
JS4-02001800-22-5A	2 – 18	30	2	2.2	2.5:1	2.5:1	5	200
JS4-02001800-50-5A	2 – 18	30	2	5	2.5:1	2.5:1	5	200
JS3-02002600-33-5A	2 – 26	21	2.5	3.3	2.5:1	2.5:1	5	150
JS3-02002600-40-5A	2 – 26	21	2.5	4	2.5:1	2.5:1	5	150
JS3-06001800-16-5A	6 – 18	23	1.8	1.6	2:1	2:1	5	125
JS3-06001800-30-5A	6 – 18	23	1.8	3	2:1	2:1	5	125
JS4-06001800-145-5A	6 – 18	31	2	1.45	2:1	2:1	5	200
JS4-06001800-30-5A	6 – 18	31	2	3	2:1	2:1	5	200

Actual
18 to 40 GHz Design

MITEQ's JS SERIES AMPLIFIERS

- High Performance/Price Ratio
- Superior, Rugged Technology
- Low Phase Distortion Design

MODEL NUMBER	FREQUENCY RANGE (GHz)	GAIN (dB, Min.)	GAIN VARIATION (±dB, Max.)	NOISE FIGURE (dB, Max.)	VSWR IN	VSWR OUT	POWER OUT @ 1 dB COMP. (dBm, Min.)	DC POWER @ +15 V (mA, Nom.)
MULTIOCTAVE BAND AMPLIFIERS (continued)								
JS3-08001800-16-5A	8 – 18	24	1.5	1.6	2:1	2:1	5	150
JS3-08001800-30-5A	8 – 18	24	1.5	3	2:1	2:1	5	150
JS4-08001800-145-5A	8 – 18	32	2	1.45	2:1	2:1	5	200
JS4-08001800-30-5A	8 – 18	32	2	3	2:1	2:1	5	200
JS3-12002600-25-5A	12 – 26	22	2.5	2.5	2.2:1	2.2:1	5	150
JS3-12002600-35-5A	12 – 26	22	2.5	3.5	2.2:1	2.2:1	5	150
JS4-12002600-22-5A	12 – 26	32	2.2	2.2	2:1	2:1	5	200
JS4-12002600-35-5A	12 – 26	32	2.2	3.5	2:1	2:1	5	200
JS3-18004000-38-5A	18 – 40	16	2.5	3.8	2.5:1	2.5:1	5	150
JS3-18004000-50-5A	18 – 40	16	2.5	5	2.5:1	2.5:1	5	150
JS4-18004000-30-5A	18 – 40	23	2.5	3	2.5:1	2.5:1	5	200
JS4-18004000-50-5A	18 – 40	23	2.5	5	2.5:1	2.5:1	5	200
ULTRAWIDE BAND AMPLIFIERS								
JS2-00100200-07-5A	0.1 – 2	32	1	0.7	2:1	2:1	5	295
JS2-00100200-15-5A	0.1 – 2	32	1	1.5	2:1	2:1	5	295
JS2-00100400-08-5A	0.1 – 4	27	1	0.8	2:1	2:1	5	200
JS2-00100400-12-5A	0.1 – 4	27	1	1.2	2:1	2:1	5	200
JS2-00100600-10-3A	0.1 – 6	23	1.5	1	2:1	2:1	3	175
JS2-00100600-20-3A	0.1 – 6	23	1.5	2	2:1	2:1	3	175
JS2-00100800-13-0A	0.1 – 8	20	1.5	1.3	2:1	2:1	0	175
JS2-00100800-25-0A	0.1 – 8	20	1.5	2.5	2:1	2:1	0	175
JS3-00101000-20-5A	0.1 – 10	23	1.5	2.0	2.5:1	2:1	5	150
JS3-00101000-35-5A	0.1 – 10	23	1.5	3.5	2.5:1	2:1	5	150
JS3-00101200-21-5A	0.1 – 12	23	1.5	2.1	2.5:1	2:1	5	150
JS3-00101200-35-5A	0.1 – 12	23	1.5	3.5	2.5:1	2:1	5	150
JS3-00101800-24-5A	0.1 – 18	23	1.8	2.4	2.5:1	2.2:1	5	150
JS3-00101800-40-5A	0.1 – 18	23	1.8	4	2.5:1	2.2:1	5	150
JS4-00101800-23-5A	0.1 – 18	29	1.8	2.3	2.5:1	2.2:1	5	200
JS4-00101800-40-5A	0.1 – 18	29	1.8	4	2.5:1	2.2:1	5	200
JS4-00102000-25-5A	0.1 – 20	28	1.8	2.5	2.5:1	2.5:1	5	200
JS4-00102000-35-5A	0.1 – 20	28	1.8	3.5	2.5:1	2.5:1	5	200
JS3-00102600-33-5A	0.1 – 26	20	2.5	3.3	2.5:1	2.5:1	5	150
JS3-00102600-42-5A	0.1 – 26	20	2.5	4.2	2.5:1	2.5:1	5	150
JS4-00102600-28-5A	0.1 – 26	27	2.5	2.8	2.5:1	2.5:1	5	200
JS4-00102600-50-5A	0.1 – 26	27	2.5	5	2.5:1	2.5:1	5	200
JS4-00104000-65-5A	0.1 – 40	14	4.5	6.5	2.75:1	2.75:1	5	200
JS4-00104000-85-5A	0.1 – 40	14	4.5	8.5	2.75:1	2.75:1	5	200

For additional information or technical support, please contact either Rosalie DeSousa at (631) 439-9458, e-mail rdesousa@miteq.com or Rizwan Syed at (631) 439-9267, e-mail rsyed@miteq.com

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for the US Navy." He will point out how GFCIs are used in underwater repairs, on airfields, and in other wet environments that pose severe challenges to performance. Citing studies that have shown units designed to Underwriters Laboratories (UL) standards to be inherently unsafe when used in these severe environments, Kampmeyer, emphasizes that design efforts should be directed at achieving the highest levels of reliability for fast-reacting GFCI hardware.

In the CAE track, Stewart Elder of Protosolv Corp. will offer relief for designers of system-on-a-chip (SoC) devices. In his talk, "Rapid Prototyping for System on Chip Verification," he will inform SoC designers how they can increase hardware and software verification performance far beyond current levels and reduce verification costs to a fraction of the current levels. Another CAE presentation, "Electronic Circuit Simulation Using the High-Level Architecture," by John Fay and Theodore Holoway of Sverdrup Technology's TEAC Group, describes the design, operation, and performance of personal-computer (PC)-based C++ software which uses the DoD High-Level Architecture (HLA) to emulate the performance of generic MC68HC11 microcontroller-based electronic circuits with near real-time response. Mohamed Bendame of MathSoft will also cover CAE, with his presentation "Design Engineering Calculations: The Hidden Asset (and Liability) in Your Product Development Life-Cycle."

In a session on simulators, Randal Burnette of Synergent Technologies will discuss the use of COTS test equipment for radar-emitter simulation, in his presentation "Radar Emitter Simulation Using Vector Signal Generators." This double-length presentation will show how to use COTS vector-signal generators to create the complex waveforms needed for radar-system-design verification.

Finally, Doug Bailey of Chip Express, in a session on ASPs and DSPs, will describe the steps that must be taken to recreate an IC that may have become obsolete, in his presentation "Resurrecting An Obsolete ASIC."

In addition to these technical sessions, the 2002 MES offers a compact but comprehensive exhibition hall. Exhibitors include such companies as Agilent Technologies, American Microwave, Ansoft, Avnet, Boonton, Elcom Technologies, Endwave, International Rectifier, M/A-COM, NoiseCom, Northrop Grumman, Raytheon RF Components, Synergy Microwave, Tektronix, Tensolite, Times Microwave, and T-Tech. For more information on the 2002 MES, please visit the website at www.mes2002.com. MRF

Antenna Offers 550-KW Peak Power Handling

MODEL 0132-800 IS a C-Band weather radar antenna that operates over the 5.25-to-5.35-GHz frequency band. The prime-focus antenna is circular paraboloid with orthogonal linear polarization. Gain is 43 dBi minimum with a 1.2-deg. maximum 3-dB beamwidth. Sidelobes are 27 dB maximum and bandwidth coincidence is 0.08 deg. maximum with polarizations. Power handling is 550 KW peak with a 1-percent duty cycle and VSWR is 1.2:1 maximum. Reflector diameter is 4.5-M nominal with two CPR-159 flanges at the reflector's edge. Pressurization is 7.5-psig desiccated air through feeders.

Seavey Engineering Associates, Inc., 28 Riverside Dr., Pembroke, MA 02359; (781) 829-4740, FAX: (781) 829-4590, Internet: www.seaveyantenna.com.

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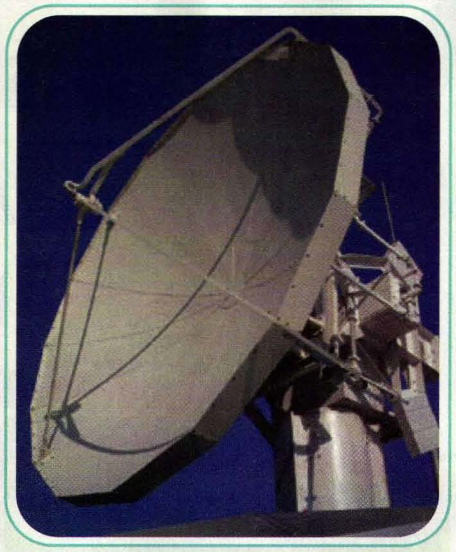

SEAVEY ENGINEERING
MODEL 0132-800

Filter Targets C-Band Applications

MODEL 7893D is a low-profile bandpass filter that is designed specifically for the C-Band market. Typical insertion loss is less than 0.4 dB at center frequency with 0.5-dB roll-off at the band edges. Rejection of 25 dB minimum at 3.65 and 4.25 GHz provides unequaled suppression of marine and altimeter interference frequencies. In addition, the unit provides a full +75-dB rejection at 3.5 and 4.4 GHz. Group delay of less than 8 nS makes this unit suitable for digital applications. Since the unit does the filtering at the C-Band frequencies, low-noise amplifier (LNA) and low-noise-block-downconverter (LNB) overload is prevented and overall picture quality is improved. The filter is constructed with copper (Cu) waveguide.

Microwave Filter Co., 6743 Kinne St., East Syracuse, NY 13057; (800) 448-1666, (315) 438-4700, FAX: (888) 411-8860, (315) 463-1467, e-mail: mfcsales@microwavefilter.com, Internet: www.microwavefilter.com.

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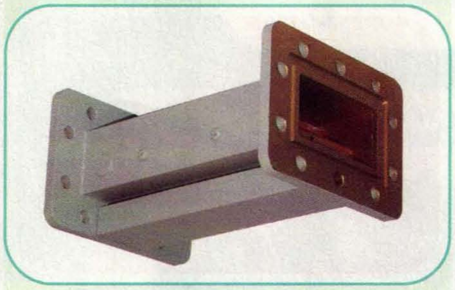

MICROWAVE FILTER CO.
MODEL 7893D

System Performs Bluetooth Rx/Tx Tests

THE ME7865A BLUETOOTH™ prequalification test system incorporates the Bluetooth-qualified MT8850A Test Set with other hardware and exclusive software to create a solution that performs power measurements on Bluetooth radios according to the RF test specification. The ME7865A addresses all 16 test cases defined in the Bluetooth RF test specification up to 3 GHz. Among the transmitter (Tx) tests that can be performed are output power, power density, power control, transmit output spectrum measurements, and modulation tests. Receiver (Rx) tests include sensitivity, carrier-to-interference performance, intermodulation (IM) performance, and maximum input level.

Anritsu Co., 1155 East Collins Blvd., Richardson, TX 75081; (800) ANRITSU, (972) 644-1777, FAX: (972) 644-3416, Internet: www.us.anritsu.com.

● **Enter No. 62 at www.mwrf.com**

ANRITSU CO.
MODEL ME7865A

Generator Tests Phase-Noise Impact

MODEL CS2040PNG is a precision phase-noise generator that is specifically designed to meet the challenges of modeling the phase noise of voltage-controlled oscillators (VCOs), synthesizers, dielectric resonator oscillators (DROs), and crystal oscillators. The generator enables the user to specify up to 16 continuous phase-noise break points and 16 spurious phase-noise values through point-and-click control. Frequency range is 1 Hz to 1 MHz and accuracy is 0.1 dBc/Hz. The device feature integrated and spurious phase-generation capability. This device replaces traditional computer modeling of performance with verifiable measurements of phase-noise-level impact and distribution.

Celerity Systems, Inc., 10411 Bubb Rd., Cupertino, CA 95014; (408) 873-1001, Internet: www.celeritydbt.com.

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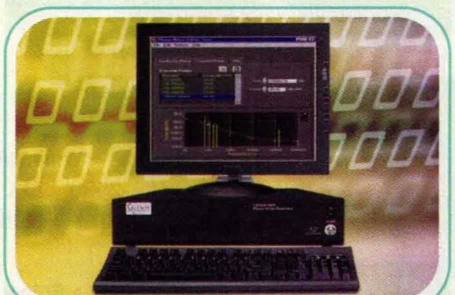
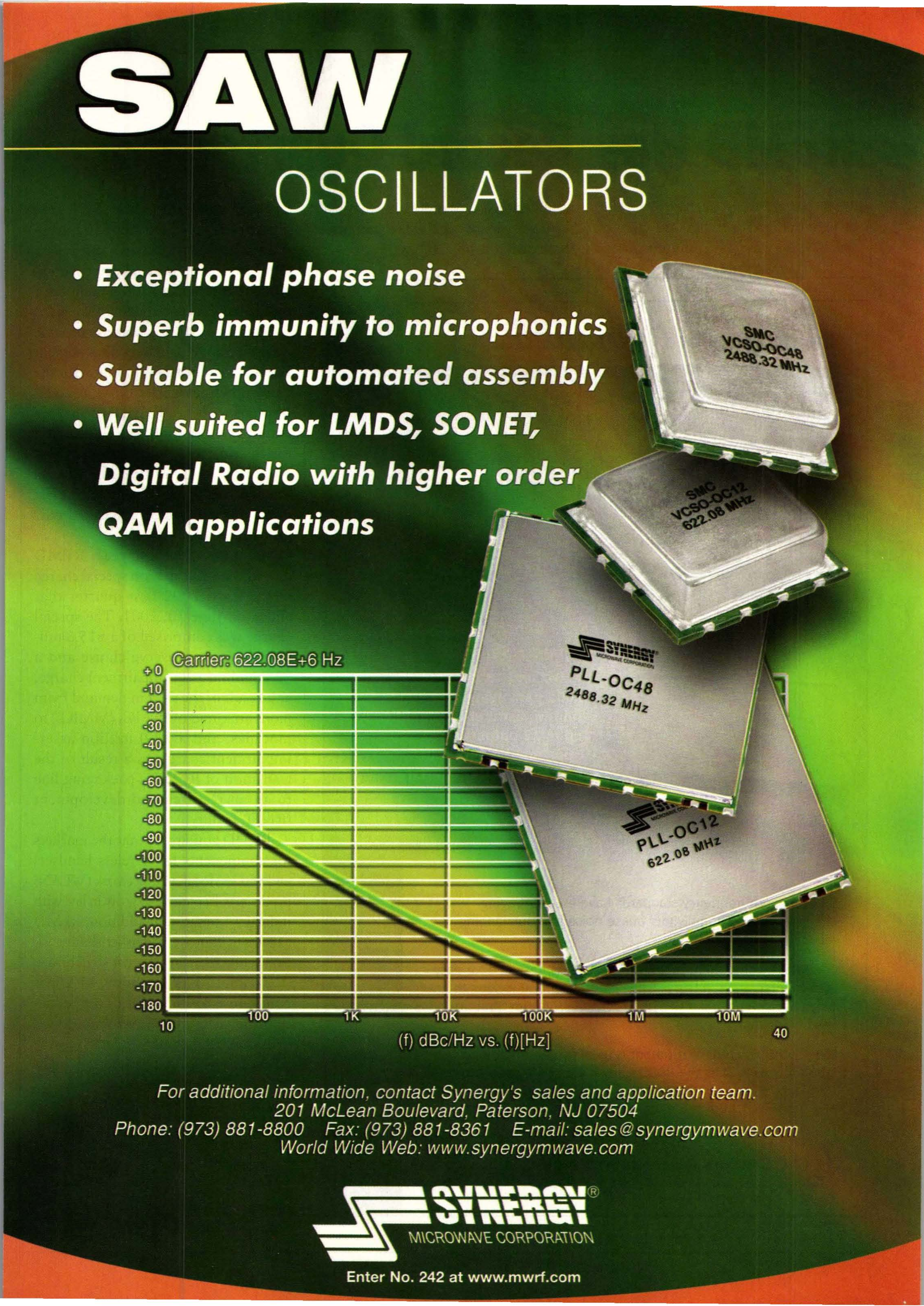


CELERITY SYSTEMS
MODEL CS2040PNG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OSCILLATORS
• Exceptional phase noise
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• Suitable for automated assembly
• Well suited for LMDS, SONET, Digital Radio with higher order QAM applications
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SMC
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622.08 MHz
SYNERGY
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2488.32 MHz
PLL-OC12
622.08 MHz
Carrier: 622.08E+6 Hz
(f) dBc/Hz vs. (f)[Hz]
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RFMD Has Record Quarter

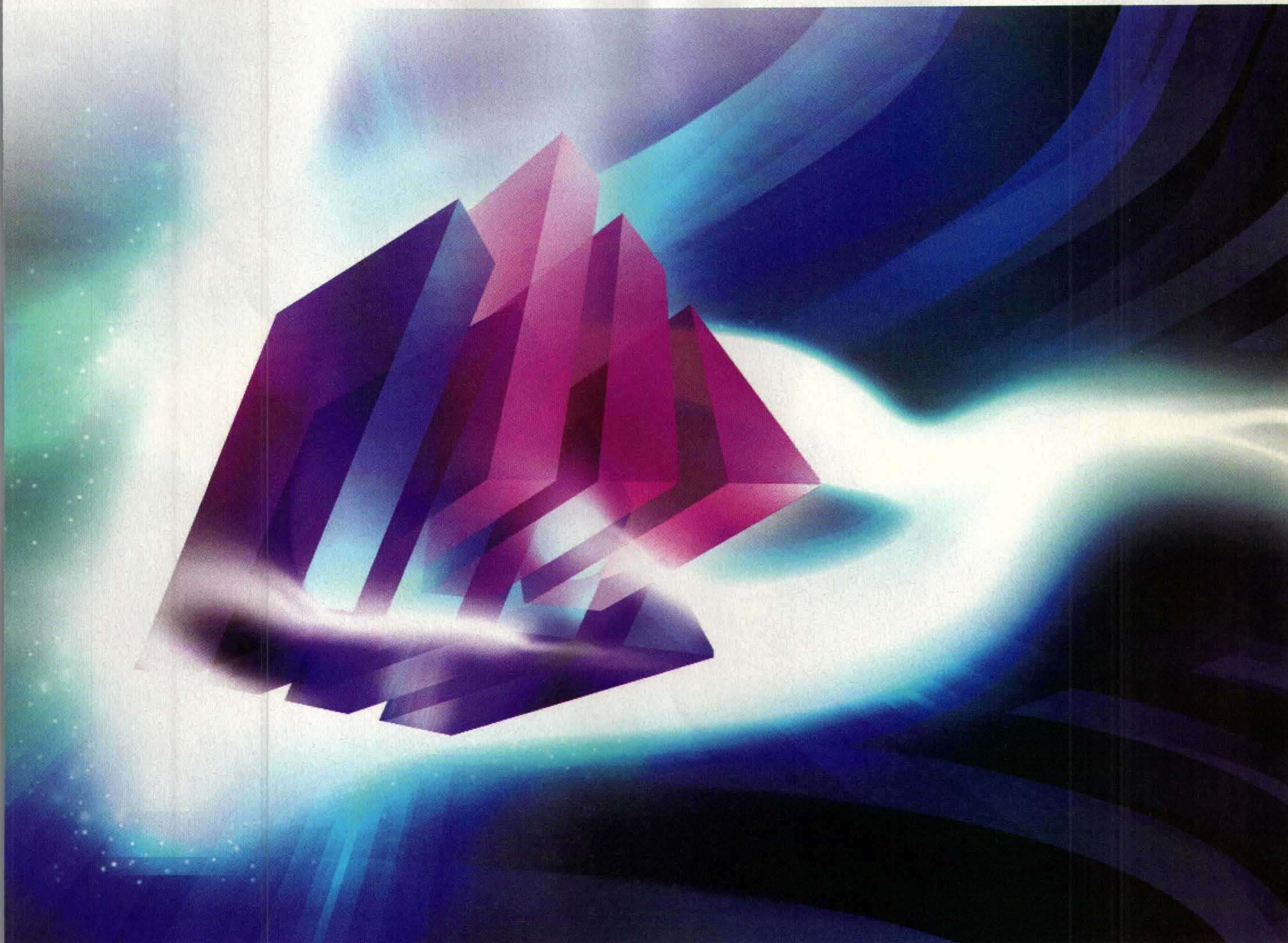
RF MICRO DEVICES, INC. (RFMD), a provider of proprietary RF integrated circuits (RF ICs) for wireless-communications applications, has reported financial results for its fiscal 2003 first quarter that ended on June 30, 2002.

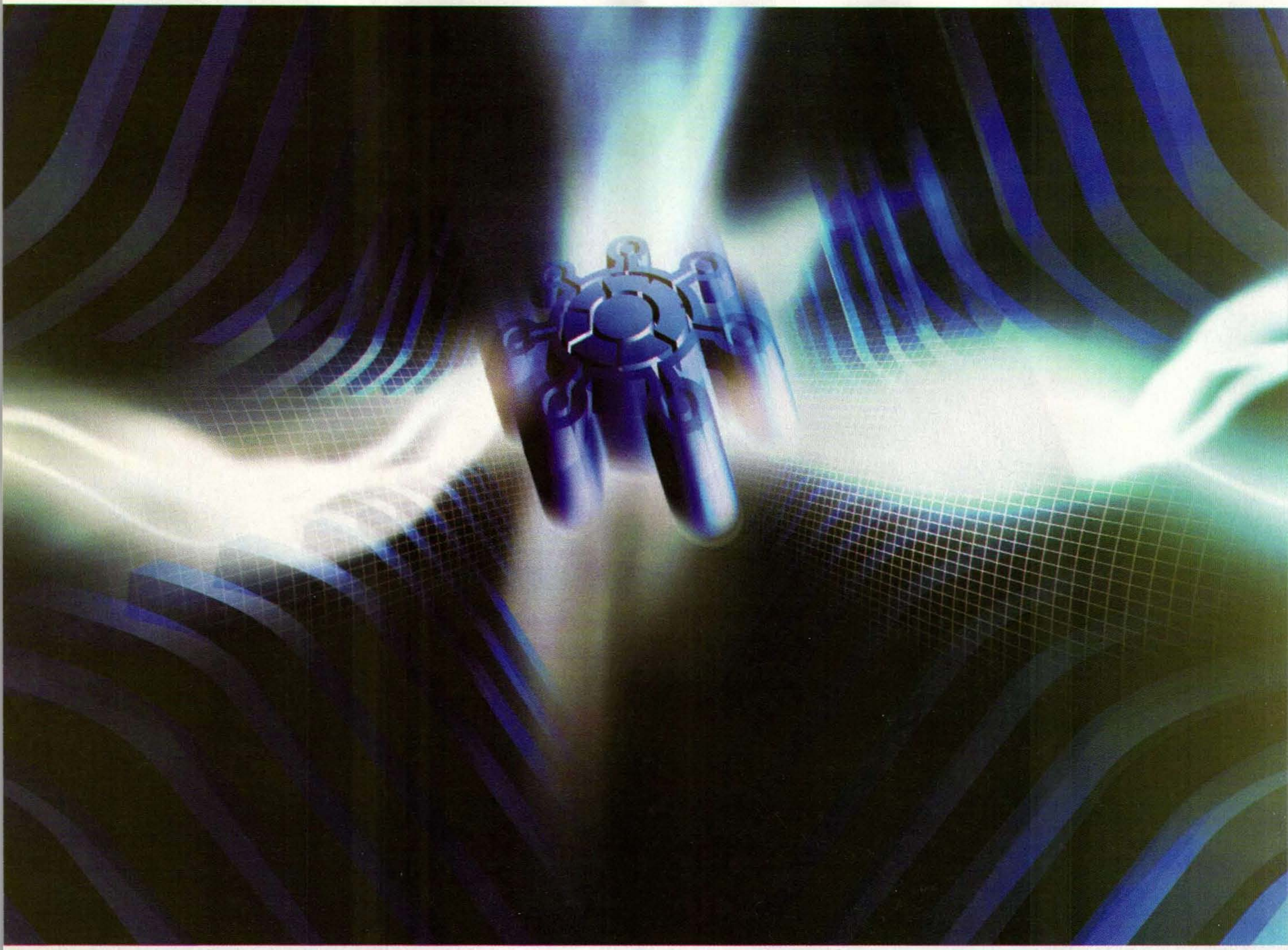
Revenue for the fiscal 2003 first quarter was approximately $103.9 million, which set a record and was a sequential increase of 3.5 percent compared to revenue of $100.4 million for the fiscal 2002 fourth quarter that ended on March 31, 2002. Quarterly revenue increased approximately 48.4 percent versus revenue of $70.1 million for the corresponding quarter of fiscal 2002. Revenue exceeded RFMD's prior estimate of $98 million to $101 million for the quarter, which was provided on June 7, 2002, because components shipped to the handset market in the last month of the quarter exceeded customer forecasts.

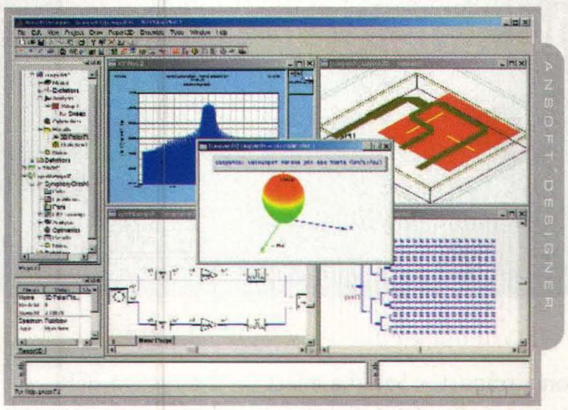
As announced last year, RFMD incurred a $22.1 million special charge in the fiscal 2002 first quarter that ended on June 30, 2001. The special charge was comprised of a $15.3 million inventory reserve clause and a $2.8 million asset-impairment charge, as a result of the shift in demand from monolithic microwave ICs (MMICs) to modules, and a $4.0 million asset-impairment charge, as a result of the transition of RFMD's packaging line to an all research-and-development (R&D) facility.

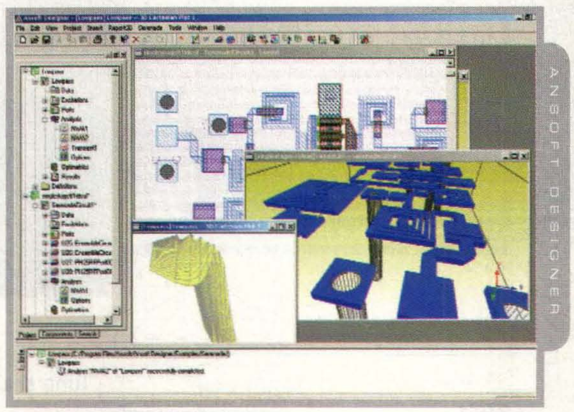
RFMD anticipates that the markets for components for wireless handsets and wireless local-area-network (WLAN) products will continue to grow in line with industry estimates. In addition, RFMD expects to improve its market-share position while reducing its cost structure.

To improve profitability, RFMD is continuing to implement cost-reduction and yield-improvement initiatives in its core handset market, including the development of next-generation, highly integrated power amplifiers (PAs). In addition, RFMD will continue investment in new markets, such as WLAN, Bluetooth™, satellite radio, Global Positioning Systems (GPS), and wireless infrastructure. Participation in these markets will likely improve margins while diversifying revenues. MRF






ANSOFT DESIGNER


ANSOFT DESIGNER

AWR

Tune the bias on a transistor. See the effect on the system BER—virtually in real time. That's the speed of the new Visual System Simulator 2002. It's from AWR, so it doesn't demand a month of training or a ton of memory to get up to speed. And the Windows interface is so intuitive you'll have the phase noise specs and amplifier linearity requirements nailed down in no time. Apply complex modulation schemes. Change encoding algorithms. Or throw in a little interference. VSS tells you the impact quickly, efficiently and—most important—accurately. Whether you're simulating an EDGE base station, 802.11 RFIC or OC-768 network, VSS lets you get the high-level perspective while keeping the wheels firmly on the ground. Download a 30-day evaluation from www.appwave.com or call us at 310-726-3000 for more information.

Visual System VSS Simulator 2002

CONTRACTS

TECOM Industries—Have received a contract award from General Dynamics Advanced Information Systems (GD-AIS) in Mountain View, CA.

TECOM will provide Crossed Polarized Log Periodic Antenna Systems and High Frequency Monopole antennas to be used in surveillance and intelligence-gathering applications.

EMS Technologies, Inc.—Has received authorization to begin work on key antenna components for Lot 11 of the Joint Army/Air Force Joint Surveillance Target Attack Radar Systems (JSTARS). The program value to EMS is approximately $1.6 million.

EMS's Space & Technology Group will manufacture and deliver phase-control components to Northrop Grumman Corp., Norden Systems. These components are necessary for the JSTARS radar and its ability to accurately map the battlefield and track targets.

Norden Systems originally contracted with EMS to design and fabricate 1400 phase shifters and 1000 twist adapters in support of the JSTARS Full Scale Engineering Development (FSED) program. Since that time, EMS has fabricated more than 6000 components.

FRESH STARTS

Sirenza Microdevices—Has established a new sales office in Stockholm, Sweden and appointed Sam Taban as the company's strategic accounts director for Sweden and Finland.

Applied Wave Research, Inc. (AWR) and CMEA Ventures (CMEA)—Announced the completion of AWR's Series B Preferred Stock financing totaling $7.4 million. CMEA, who will be taking an equity position and providing representation on AWR's board of directors, led the round. The investment will be used to accelerate the development of the company's high-frequency electronic-design-automation (EDA) solutions and expand AWR's worldwide sales organization.

Superconductor Technologies, Inc.—Announced financial results for the quarter and six months ended June 29, 2002.

Total net revenues for the second quarter of 2002 were $6.1 million, an increase of 44 percent versus $4.2 million for the second quarter ended June 30, 2001, and an increase of 31 percent over the first quarter of 2002. Net commercial product revenues for the second quarter of 2002 were $5.3 million, an increase of 81 percent compared to $2.9 million in the year-ago period, and an increase of 42 percent over the first quarter of 2002. Government contract revenues for the second quarter of 2002 were $785,000 compared to $1.3 million in the second quarter that ended on June 30, 2001.

The total net loss for the quarter that ended on June 29, 2002 was $6.0 million versus a loss of $3.5 million in the quarter that ended on June 30, 2001. The net loss available to common stockholders for the second quarter of 2002 was $6.6 million, or $0.29 per diluted share, compared with the loss of $4.1 million, or $0.23 per diluted share for the year-ago quarter that ended on June 30, 2001.

QUALCOMM, Inc.—Revealed its third-quarter fiscal 2002 results ending on June 30, 2002. Pro forma revenues were $721 million, an increase of 10 percent compared to $657 million in the year-ago quarter and 9 percent compared to $659 million in the second quarter of fiscal 2002. Pro forma earnings per share were $0.24 in the third quarter of fiscal 2002, an increase of 20 percent compared to $0.20 per share in the year-ago quarter and the second quarter of fiscal 2002. GAAP reported revenues for the third quarter of fiscal 2002 were $771 million compared to $657 million in the year-ago quarter and $696 million in the second quarter of fiscal 2002. GAAP reported loss was $14 million or $0.02 per share in the third quarter compared to a loss per share of $0.26 in the year-ago quarter and earnings per share of $0.05 in the second quarter of fiscal 2002.

Tech-Etch, Inc.—Acquired certain assets of BMC Industries, Inc.'s Buckbee-Mears Group's Micro-Technology Operation associated with their sheet-etching business located in St. Paul, MN. These assets will be relocated to Tech-Etch's photo-etching business in Plymouth, MA. The technology, equipment, and customer business acquired will combine with Tech-Etch's position in the precision-parts photo-etching business serving the medical, electronic, and fuel-cell markets. Buckabee-Mears' strength in the etching of exotic materials such as Tungsten (W), Molybdenum (Mo), Titanium (Ti), and Elgiloy will enhance Tech-Etch's capabilities to serve these markets.

Verint Systems, Inc.—Announced that its LORONIX video solution has been implemented in the Mall of America in Bloomington, MN to help protect the safety and security of its 42.5 million annual visitors and 12,000-plus employees. The Mall of America is the largest enclosed retail and entertainment complex in the US. The LORONIX solution provides a networked digital security system that covers the entire facility, including the indoor amusement park.

Verint's LORONIX system, which monitors the Mall of America's 4.2 million sq. ft. of interior space plus parking lots, decks, and other common areas of the 75-acre facility, allows for all cameras to be viewed and operated from one centralized control room. The LORONIX solution's automated alarms, which are based on video-content analysis, are designed to make surveillance more effective and to free security staff for other activity. The LORONIX solution permits the use of additional cameras without requiring additional security personnel to monitor the system. This permits the Mall of America to keep its staffing expenses under control while maintaining a high level of security and protection. MRF

GROSSMAN

Grossman Named President And CEO At e-tenna Corp.

STEVEN GROSSMAN has been appointed as president and CEO of e-tenna Corp. He formerly served as senior vice president and general manager of the Flash Memory Business Unit at Hynix Semiconductor (formerly Hyundai Electronics).

Cushcraft Corp.—GREGORY S. CZUBA to president and COO; formerly president of REMEC Wireless.

Concierge Technologies, Inc.—DAVID W. NIEBERT to president and COO; remains as president of The Wallen Group.

Tropian, Inc.—HOREN CHEN, PH.D. to president and CEO; previously founded Mobilink Telecom.

EMS Technologies, Inc.—RON H. MIYAKAWA to vice president of business development; formerly operations manager, Commercial RF, with direct responsibilities as the campaign manager for the Anik F2 satellite program at Boeing Satellite Systems.

Rockwell Collins—D.W. (DENNIS) HELGESON to the position of vice president and general manager of Business and Regional Systems for Commercial Systems; formerly vice president for business development with the Government Systems division.

Kyocera Wireless Corp.—HOWARD W. "SKIP" SPEAKS to the position of CEO; remains as president.

Danaher Corp.—KURT GALLO to the position of president of Danaher Power Solutions; remains as vice president and general manager of the Enterprise Group.

KVH Industries—PATRICK J. SPRATT to CFO; formerly CFO and treasurer of BioReliance Corp. Also, RICHARD FORSYTH to vice president for finance; formerly CFO.

JMAR Technologies, Inc.—JOSEPH G. MARTINEZ to interim CEO; formerly held the position of senior vice president and general counsel.

RF Micro Devices—DR. ALBERT E. PALADINO to chairman of the board of directors; formerly outside director.

Photofabrication Engineering, Inc.—KIRK LAUVER to staff engineer responsible for process improvements and optimization; formerly product manager at Atotech USA.

The National Electrical Manufacturers Association (NEMA)—KENNETH HONEYCUTT to the Board of Governors; remains as president and CEO of Acuity Lighting Group. Also, GREGORY B. KENNY to the Board of Governors; continues as president and CEO of General Cable. In addition, THOMAS N. MCCAUSLAND to the Board of Governors; remains as president and CEO of Siemens Medical Solutions. And, JAMES L. PACKARD to the Board of Governors; continues as chairman and CEO of Regal-Beloit.

International Wafer Service, Inc. (IWS)—GREG TEN EYCK to Microfab technical sales engineer; formerly was a co-founder and director of Haleos/ACT MicroDevices.

TEN EYCK

FERGUSON

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Examine An HTS Dual-Mode, Cross-Slotted Filter

FILTERS AND ANTENNAS that are used as part of a high-temperature superconductor (HTS) microwave device show considerably improved performance, lower power dissipation, and reduced size in telecommunications systems over their traditional counterparts. However, these improved properties tend to degrade as power increases due to the nonlinear nature of the HTS surface resistance, which increases with the circulating RF current.

As such, a compromise between component size and power must be reached in an attempt to stave off this degradation. The result of this necessary compromise has been the design of dual-mode, cross-slotted C-Band filters that present compact size and Chebyshev or quasi-elliptical responses, offering widely spread current distribution and relatively good power handling. These filters have been grown on lanthanum-aluminum-oxide ($LaAlO_3$) substrates. See "Superconducting Dual-Mode Dual-Stage Cross-Slotted Filters," *Microwave And Optical Technology Letters*, Vol. 33, No. 6, pp. 389-392.

Design A Chip Antenna For GSM/DCS Mobile Phones

CHIP ANTENNAS ARE most often manufactured using ceramic chips. Relative permittivity for these antennas is usually large (approximately 7.0), making wide impedance bandwidth difficult to achieve when a grounded conducting plane is printed on the bottom of the ceramic chip. However, it has been discovered that when a rectangular spiral conducting strip is placed on the bottom, top, and two side surfaces of a rectangular plastic chip measuring $5 \times 10 \times 16$ mm^2, a plastic chip antenna that is capable of Global System for Mobile Communications/digital communications services (GSM/DCS) dual-band operations in mobiles phones can be created. This antenna has the bonus of being surface-mountable and providing good antenna performance. See "Dual-Band Plastic Chip Antenna for GSM/DCS Mobile Phones," *Microwave And Optical Technology Letters*, Vol. 33, No. 5, pp. 330-332.

Maintain Transition Frequency With Low Noise Using A 7-GHz LNA

DUE TO RECENT improvements to complementary-metal-oxide-semiconductor (CMOS) technology, a transition frequency f_T can now be viewed as competing with silicon-bipolar-junction-transistor (Si-BJT) technology. Thermal noise of base resistance and shot noise of the collector current are specified as the main noise sources for Si-BJTs. Reduction of this noise requires a large geometry device that uses more collector current to maintain high f_T value, but which causes more shot noise. To solve this conundrum, researchers have developed a 7-GHz low-noise amplifier (LNA) that is fabricated using 0.25-μm CMOS technology. A cascode configuration with a dual-gate MOSFET and shielded pads were adopted to improve the gain and the noise performance. An associated gain of 8.9 dB, a minimum noise figure of 1.8 dB, and an input-referred third-order intercept point (IP3) of +8.4 dBm were obtained at 7 GHz. The LNA consumes 6.9 mA from a +2-VDC supply voltage. These results indicate the capability of a CMOS LNA using these techniques for low-noise and high-linearity applications at more than 5 GHz. See "A 7-GHz 1.8-dB NF CMOS Low-Noise Amplifier," *IEEE Journal of Solid-State Circuits*, Vol. 37, No. 7, pp. 852-856.

Reduce SNR Degradation In EM Environments Using A Nulling Technique

ARRAY-PATTERN NULLING effects have become an important field of study recently due to the increased pollution of electromagnetic (EM) environments. These techniques reduce degradation of signal-to-noise-ratio (SNR) performance due to undesired interference in radar, sonar, and communications systems. Researchers have worked with a flexible nulling method based on the ant-colony-optimization algorithm for the pattern synthesis of linear antenna arrays. Pattern nulling is achieved by controlling on the amplitude of each array element. Researchers show the versatility of the method by considering some design specifications such as sidelobe level, null depth, and dynamic-range ratio and introducing a set of weighting factors in the cost function constructed for the ant-colony-optimization algorithm. Several examples of Chebyshev patterns with imposed nulls are provided. See "Null Steering of Linear Antenna Arrays with Use of Modified Touring Ant Colony Optimization Algorithm," *International Journal of RF and Microwave Computer-Aided Engineering*, Vol. 12, No. 4, pp. 375-383.

DESIGN

Efficiently Modulate UWB Signals

UWB systems can employ power-efficient modulation to achieve its promise of high wireless data rates with relatively low average transmit power.

Ultrawideband (UWB) technology has been described by some as one of the most promising technologies of our times. Early UWB systems were developed mainly as a military surveillance tool due to their ability to "see through" trees and beneath ground surfaces. But recently, UWB technology has been focused toward the area of commercial communications. This article explores some of the reasons why UWB differs from conventional narrowband radios and the importance of modulation selection when designing UWB communication systems.

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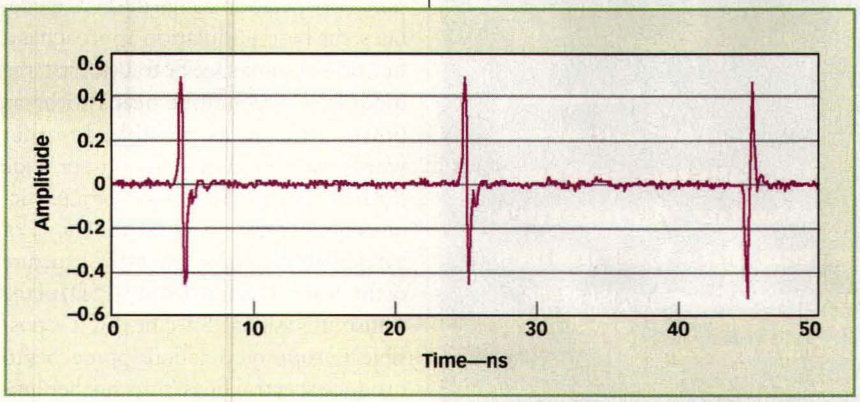
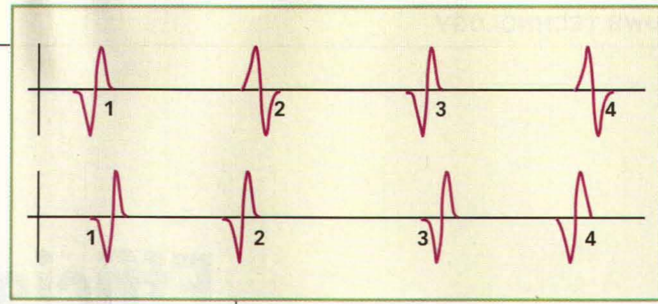
As their name implies, UWB systems operate across a wide range of frequency spectrum. One typical form of UWB signal is created by the transmission of a series of low-power pulses. A very simple example of this signal is shown in **Fig. 1.** The UWB pulses are extremely short in duration (10 to 1000 ps), typically much shorter than the time interval corresponding to a single bit in the transmitted data stream. This is similar in nature to the unintentional emissions referred to as "noise" generated by all-digital devices (computers, televisions, answering machines, etc.). Indeed, the transmit power for a device compliant with Federal Communications Commission (FCC) UWB regulations is less than what a standard personal computer (PC) is allowed to radiate unintentionally.



1. A train of Gaussian pulses follows a repeating amplitude pattern with time.

Due to the short duration of the pulses, the frequency spectrum of a UWB signal can be several gigahertz wide, overlaying the bands used by existing narrowband systems. This "overlaid" operation is possible because the low transmission power of a UWB signal is spread over much wider bandwidth than any conventional narrowband systems. For example, a 6-GHz-wide UWB transmitter (Tx) distributes its energy over the equivalent of 1000 television channels, 30,000 frequency-modulated (FM) radio channels, or a-half-million walkie-talkie channels, making the UWB signal power in any single narrow frequency channel extremely small. This fact is a primary basis

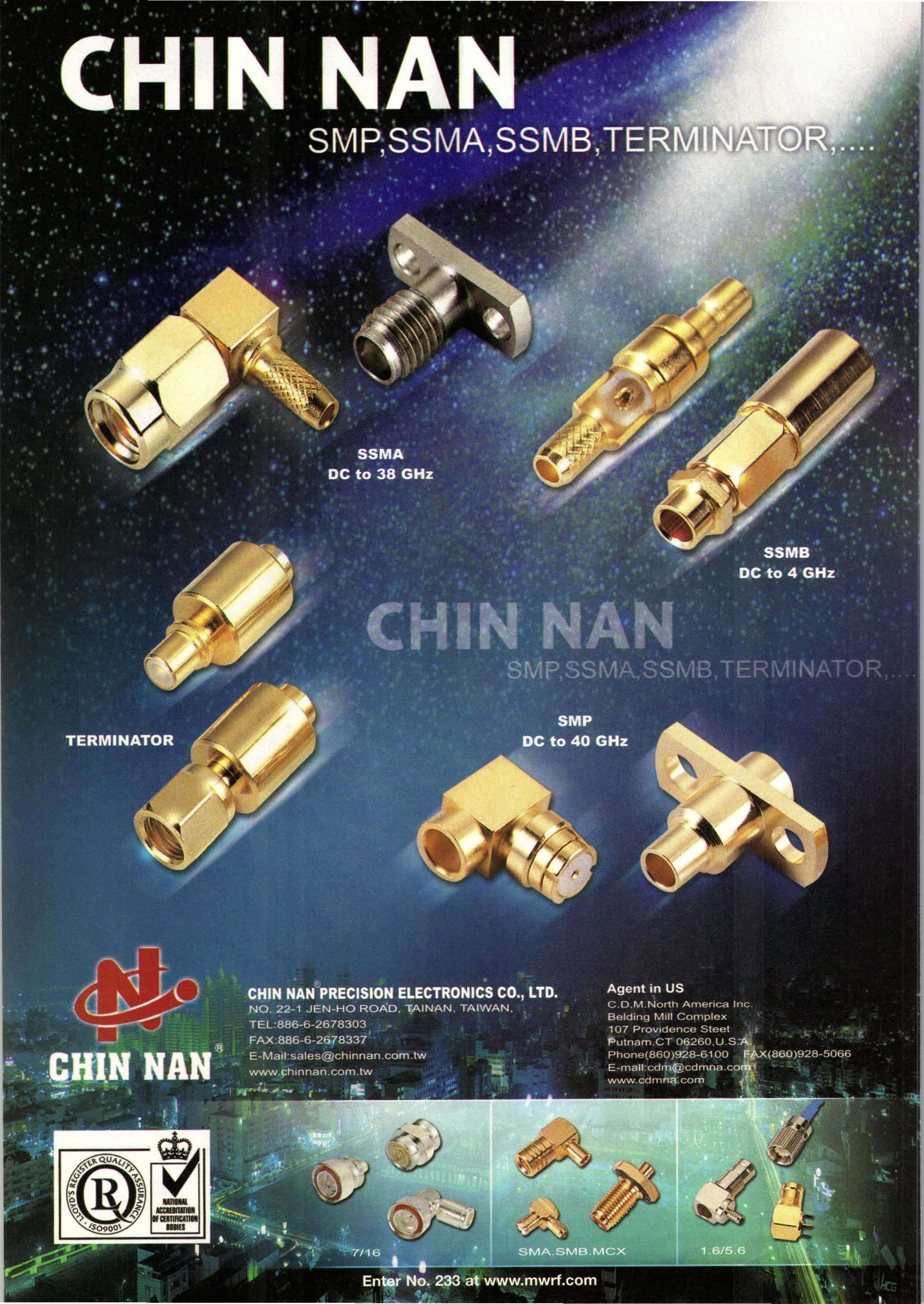


2. These pulse trains show BPSK (top) and pulse-position modulation [PPM] (bottom) as a function of time.

for the recent FCC decision to allow the commercial sale and usage of UWB devices within defined frequency and power limits. The FCC emission limits are equivalent to a transmission level of less than 75 nW of power per megahertz of bandwidth between the frequencies of 3.1 and 10.6 GHz. This roughly equates to a total aggregate output power of 0.5 mW—significantly less than the 30 to 100 mW permitted for other unlicensed devices such as IEEE 802.11b wireless local-area-network (WLAN) radios.

From a communications-theory perspective, perhaps the most important characteristic of UWB systems is their power-limited regimen of operation. This is in contrast to many conventional radio systems that operate in a bandwidth-limited regimen where, to obtain more data rate, a sacrifice must be made in either signal processing (i.e., more complex circuitry leading to additional silicon (Si) chip area being employed) or transmit power. For instance, IEEE 802.11a or HiperLAN2 systems use a 64QAM modulation when transmitting at 54 Mb/s in a 25-MHz channel. In this case a relatively high bits-energy-to-noise-power-density (E_b/N_o) ratio is needed to effectively demodulate the higher-order modulation, so the system must be operated with lower power efficiency to achieve higher spectral efficiency. In contrast, UWB systems are not as limited in bandwidth, so they can use more power-efficient modulation techniques to achieve high data rates with relatively low transmit power.

To create UWB communications signals, there are a number of different methods that can be used for modulating a pulse train with data for transmission. For operation within the FCC guidelines, the best modulation approach is a function of some specific trades. First, the modulation technique needs to be as power efficient as possible. In other words, the modulation needs to provide the best error performance for a particular energy per bit. Second, the choice of a modulation scheme affects the structure of the power spectral density (PSD) of the transmitted signal. Specifically, it is possible for some modulation approaches to produce spectral lines, thus further limiting total transmit power to stay with-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in FCC limits on signal PSD, leading to a lower average power solution.

Two of the most common modulation techniques that have been proposed for UWB are mono-phase techniques such as pulse-position modulation (PPM) and biphase modulation techniques such as binary phase-shift keying (BPSK). Examples of these two modulation techniques are shown in **Fig. 2** for two UWB pulse sequences.

To appreciate the impact of the modulation choice on the power efficiency of the system, consider **Fig. 3**, which shows a simple representation of the symbol constellations for the two modulation techniques. Here we assume that for the PPM system the time offsets for the pulses are chosen to make the two possible pulses orthogonal at the receiver (Rx), meaning that they do not overlap each other. This results in an orthogonal signaling scheme.

The BPSK technique is an antipodal signaling scheme and has a greater intersymbol distance than PPM for equal energy per bit. This difference leads to a 3-dB advantage in power efficiency, meaning that PPM requires twice as much transmit power as BPSK to achieve the same bit-error rate (BER) in equivalent conditions.

As mentioned previously, a general UWB pulse-train signal can be represented as a sum of pulses shifted in time:

$$s(t) = \sum_{k=-\infty}^{+\infty} a_k p(t - t_k) \qquad (1)$$

where:

s(t) = the pulse train signal,

p(t) = the basic UWB pulse, and

a_k and t_k = the amplitude and time offset of each individual pulse, respectively.

The PSD of this signal, $\Phi_{ss}(f)$, is the Fourier transform of the signal autocorrelation. If we assume that the pulse weights correspond to the data bits to be transmitted and the data is random and pulses are uniformly spaced (i.e., t_k=kT), we can find that the PSD is:

SEE EQ. 2 ON P. 65

where:

σ^2_a and μ_a = the variance and the mean of the weight sequences, respectively,

P(f) = the Fourier transform of p(t), and δ(f) = a unit impulse.

As can be seen, this PSD has a continuous portion and discrete spectral lines. It is important to note that the magnitude of the spectral lines depends on the mean of the weights, μ_a. If we wish to encode the data bits using BPSK (Fig. 1), then we use $a_k \in \{-1,+1\}$ and so σ^2_a = 1 and μ_a = 0, assuming that the data bits are random and equi-probable,

$$\Phi_{ss-PAM}(f)=\frac{\sigma_a^2}{T}|P(f)|^2+\frac{\mu_a^2}{T^2}\sum_{k=-\infty}^{\infty}\left|P\left(\frac{k}{T}\right)\right|^2\delta\left(f-\frac{k}{T}\right) \quad (2)$$

$$s(t)=\sum_{k=-\infty}^{+\infty}p(t-kT+a_k\beta T) \quad (4)$$

$$\Phi_{ss-PPM}(f)=\frac{\sigma_a^2}{T}|B(f)|^2+\frac{\mu_a^2}{T^2}\sum_{k=-\infty}^{\infty}\left|M\left(\frac{k}{T}\right)\right|^2\delta\left(f-\frac{k}{T}\right) \quad (5)$$

/hich is usually the case when employ-ng a data whitener. In this case, we n notice that the discrete portion of : equation disappears (since $\mu_a = 0$). ւ. 2 becomes:

$$\Phi_{ss-BPSK}(f)=\frac{1}{T}|P(f)|^2 \quad (3)$$

As mentioned above, it is crucial for WB to minimize the presence of spec-al lines, since the transmit power of se systems must be constrained to meet : limits on power-spectral density. ne presence of spectral lines may lead a reduction in total transmit power ess the lines can be reduced in some ler way. Additionally, spectrum lines n lead to greater potential interference ith conventional radio systems.

In contrast to PAM or BPSK, a PPM stem encodes the data bits in the pulse eam by advancing or delaying indi-lual pulses in time relative to some ference. In this case, the equation for ie UWB signal becomes:

SEE EQ. 4 ABOVE

here:

$a_k \in \{-1,+1\}$ and βT is the amount f pulse advance or delay in time rela-ve to the reference (unmodulated) osition. The form of the PSD for PPM s slightly different as well, and depends ess directly on the UWB pulse shape:

SEE EQ. 5 ABOVE

/here:

B(f) and M(f) = the Fourier transforms f $b(t) = 0.5[p(t-T) - p(t+\beta T)]$ and $m(t) = 0.5[p(t-\beta T) + p(t+\beta T)]$.

This PSD expression shows that a PPM gnal contains spectral-line content ıd suffers from the limiting ffects these may cause as oted previously.

To partially reduce the effect f the spectral lines caused by PM, some form of dithering ın be used. In general, dither-g is a random process that troduces jitter to the posi-ion of the individual pulses ıccording to a pseudorandom sequence, reducing spectral lines. This is one reason that PPM is often used in conjunction with time-hopping multiple access techniques—the time-hopping provides a degree of dithering to reduce spectral-line content relative to a nonhopped PPM signal. Although dithering can limit spectral lines, it does nothing to compensate for the sub-optimal power efficiency of PPM noted earlier, and may be hard to provide without significantly increasing the level of complexity (and, by association, Si area and cost) of a UWB system.

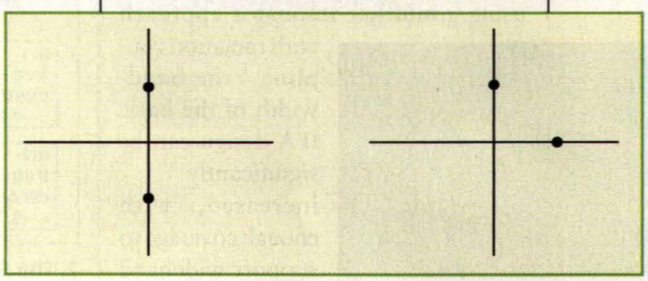
Is Biphase Better?

Prior to the FCC ruling on February 14, 2002, the main customer of UWB communication technology was the military, with requirements typically driven by longer operational ranges for lower data-rate applications (i.e., closer to 1 Mb/s than to the more than 100 Mb/s needed for consumer video applications). Most of the UWB developments for the military operated at lower frequencies (below 3 GHz) and much higher power levels than those approved by the FCC for commercial deployment. In addition, these systems typically employed monophase pulse modulation [either PPM or on-off keying (OOK)] and operated at relatively low pulse-repetition frequencies (PRF) for the underlying pulse trains.

A more effective solution for future commercial UWB systems is to combine more efficient BPSK modulation with UWB pulse trains at higher PRFs. The advantage of high PRF systems is that they can be designed to provide a low peak-power-to-average ratio (PAR), often as low as 3 or less (for reference, a sine-wave PAR is equal to 2). Consequently, an implementation using high-PRF biphase UWB pulses does not require any external snap-recovery or tunnel diodes and microwave circuitry. Due to the low PAR, the antenna can also be connected directly to the complementary-metal-oxide-semiconductor (CMOS) integrated-circuit (IC) gate switch, requiring no additional amplification and no mixers.

In addition, biphase modulation can provide reduced jitter requirements relative to low PRF PPM systems. The reason has to do with clocking. In PPM, the designer must accurately control arbitrary time positions on a fast (i.e., on a pulse-to-pulse) basis. This control requires a series of wide bandwidth circuits where jitter accumulates. By contrast, a biphase system needs only a stable clock, since the pulses occur on a constant spacing (phase modulation only).

With a CMOS IC gate-driven RF front end, the speed of the radio becomes directly a function of the speed of the IC transistor. A certain threshold transistor speed is necessary to produce a UWB waveform in the FCC-allocated band. Ten years ago, producing a biphase UWB architecture operating in the current FCC allocated band would have been cost inefficient. As IC processes have continued to produce faster transistors (as they consistently do in accordance with Moore's Law), this type of system can be produced quite affordably. As IC processes continue to increase in speed and power efficiency, even faster—or higher data rate—radios can be produced at affordable prices. CMOS-driven biphase modulation is a technique specifically well-suited for high PRF, low-power UWB – hence the Moore's Law Radio™. MRF



3. These constellation diagrams show biphase (or BPSK) modulation (left) and PPM (right).

Method Boosts Bandwidth Of IFAs For 5-GHz WLAN NICs

The use of an additional resonator and a coupled-by-radiation technique can increase the bandwidth of a traditional IFA for high-data-rate 5-GHz WLANs.

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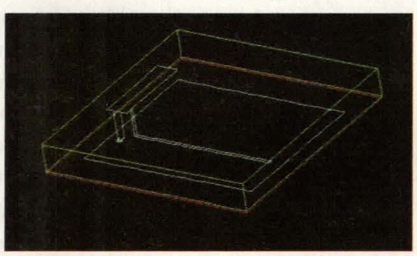
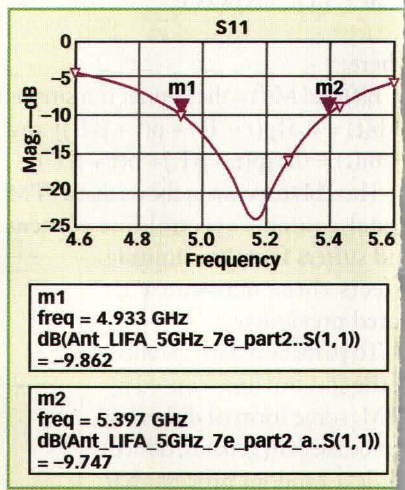
Wireless local-area networks (WLANs) at one time were associated with the unlicensed frequency band at 2.400 to 2.483 GHz, although more recent advances in WLAN technology have led to the use of the unlicensed 5-GHz band for wide-band, high-data-rate network services. In these systems, the antenna on the client network interface card (NIC) is a critical component, requiring high performance in a low-profile package but for minimum cost. The inverted F-shape antenna (IFA) is one possible antenna configuration with a great deal of promise for these 5-GHz WLAN NIC applications. The IFA features a low profile and can be stamped out in a single manufacturing step for low-cost production.

In its basic form, however, the IFA has somewhat limited bandwidth of 10 percent. Since wideband 5-GHz WLAN applications can cover a total of 700 MHz from 5.15 to 5.85 GHz, or a bandwidth of 12.7 percent at mid-band, more bandwidth is needed than the basic IFA design can deliver. By using a multiple-resonator approach with radiation coupling,[1] the bandwidth of the basic IFA design can be significantly increased, with enough coverage to support wideband 5-GHz WLAN applications.



1. This EM simulation model shows the configuration of a basic IFA.

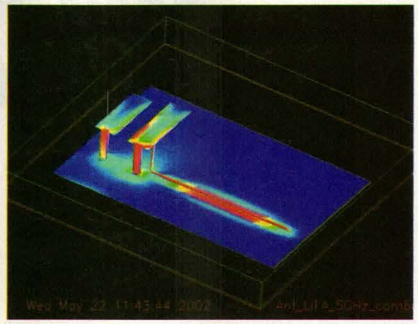
The basic IFA design ha been used in many "low-cost applications, including i Bluetooth designs, PDC minia ture Global System for Mobile Com munications (GSM) handsets, and in cod division-multiple-access (CDMA) cellula telephones. The antenna design can b easily integrated into these and other elec tronic devices. With a plastic cover, a IFA is well-concealed inside the device The plastic cover offers extra protectio



2. The S11 response as a function of fr quency was plotted for the traditional (narrowband) IFA.

for the antenna without affecting the antenna performance. The basic IFA has been well-studied and well-researched. References 1 through 7 provide more in-depth discussions on this topic.

The basic IFA is essentially a rectangularly-shaped resonator bent into an L-shape. One end is left open, while the other end is shorted to ground with a viahole connection. The RF signal is fed along the wider edge of the resonator. The feeding position can be determined experimentally. With the RF feed line on the side, the antenna structure is similar to an upside-down F, thus the name.



3. This EM simulation model shows a modified, wideband IFA, using radiation-coupled dual resonators.

The structure of a basic IFA is shown in **Fig. 1**, where a half-wavelength transmission line (TL) is used to feed the structure. The half-wave line makes de-embedding much easier since only attenuation loss must be taken into account. If the antenna structure is fed right below the antenna without the extra TL, erroneous experimental results will be observed by other modes generated due to the incorrect boundary condition.

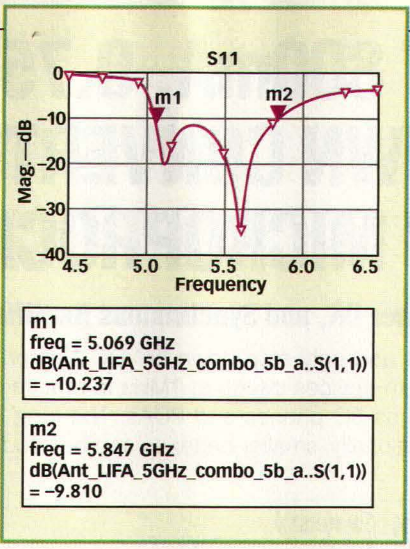
A basic IFA design is 3.5 mm high, 10.45 mm long, and 2.7 mm wide. As shown in **Fig. 2**, a basic design covers from 4.93 to 5.39 GHz, with close to 540-MHz bandwidth. While it has enough bandwidth to cover the lower portion of the 5-GHz unlicensed band, it does not provide adequate coverage (short by at least 260 MHz) to cover the entire 5-GHz unlicensed band. A possible solution for increasing the bandwidth lies in using a coupling-by-radiation technique.[1]

Using the coupled-by-radiation technique, it is possible to design a wideband antenna in the same manner as a multiresonator filter. For example, in an edge-coupled microstrip filter design, multiple resonators are placed in close proximity and, with correct spacing, a wideband bandpass-filter (BPF) response can be achieved. Similarly, a wideband antenna can be designed by adding a second resonator with correct spacing between the two resonators. One of these wideband antenna designs is shown in **Fig. 3**. In this design, the second resonator is placed 1.7 mm apart from the main resonator. The lengths of the first and second resonators are optimized to produce a resonance at the appropriate

frequency of interest **(Fig. 4)**.

This particular wideband antenna design covers from 5.07 to 5.85 GHz, with better than 10-dB return loss. It has a bandwidth that is close to 780 MHz and clearly represents an improvement over the performance of the basic IFA design. As can be seen, the depths of the two resonances are not equal. This unequal-resonance configuration tends to provide the best bandwidth performance, although the design can be tuned to provide resonances with equal depths, to match the theory presented in textbooks.

The use of electromagnetic (EM) simulation computer-aided-engineering (CAE) software was very useful in the design of this wideband IFA. In the basic IFA, the feedline position must be determined empirically. Without the EM software, finding the proper feedline position would require the construction of several prototypes and a great deal of design time. But using the EM simulator for the basic IFA design, the simulation results were found to exactly match the test data on a prototype. The EM software was a great help in determining the spacing between the two resonators without building multiple prototypes. The EM simulator in this case was the 2.5-dimensional simulation tool Momentum from Agilent Technologies (Santa Rosa, CA). MRF



4. The S11 response of the wideband IFA was plotted versus frequency.

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Gauge Nonthermal Effects Of Microwave Fields

Significant biological effects have been found even for EM field levels considered safe by global standards organizations.

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Radiation-exposure standards around the world are generally based on the induced heating effects of RF and microwave fields on biological tissues for particular field levels. But controversy continues over the possible existence of nonthermal effects of electromagnetic (EM) fields on biological materials, implying that these thermal-based standards may be based on erroneous assumptions. To better evaluate existing radiation-exposure standards, it is necessary to understand the possible nonthermal effects of RF and microwave fields on biological tissues and materials.

Thermal effects from exposure to EM fields depend on the specific-absorption-rate (SAR) distributions. The primary effect of microwave and RF energy on living tissue is an increase in temperature[1-4] due to an absorption of energy, although nonthermal effects have now also been identified.[5-13] Stress reactions due to total body exposure,[14] effects on reproduction,[15] and physiological and biological changes[16,17] in rats and humans following exposure to even low levels of RF and microwave radiation have been reported.

To understand the nonthermal effects of exposure to RF and microwave fields, two mechanisms for these nonthermal effects will be examined: the alignment of long molecules with a dipole moment, and photon and bond energy levels in biological materials. The first mechanism involves the alignment of long-chain molecules along electric-field lines. An electric field applied to a molecule with a dipole moment[18] or an induced dipole moment produces a torque which tends to align the molecule along the field line. The torque, T, is applied to the molecule by the electric field and is yielded by:

$$T = -|m|\, E \sin\theta \qquad (1)$$

where:

E = the electric field,

|m| = the magnitude of the dipole moment, and h = the angle between the dipole and E.

For a fixed dipole, |m| is given by:

$$|m| = qds \qquad (2)$$

where:

q = the charge and

ds = the separation.

In the case of an induced dipole moment, |m| is provided by:

$$|m| = |a\ E| \qquad (3)$$

where:

a = a property of the molecule.

The energy required to rotate a molecule (W) is defined by:

$$W \propto |mE| \qquad (4)$$

The threshold field for observing an orientation is dependent upon the restraining forces in the material. It has been shown[18] that the size of the currents which can result from AC electric fields are large enough that the number of charged ions or charged molecules which can be moved across a membrane are large enough to potentially cause biological changes in the periods of time.

The primary effect of microwave and RF energy on living tissue is an increase in temperature due to an absorption of energy, although nonthermal effects have now also been identified.

The second kind of nonthermal effect, a change in biological material through which it is passing, an EM wave deposit enough energy to alter some structure significantly. An EM wave consists of very small packets of energy known as photons. The energy in each photon is proportional to the frequency and it is defined as, hf, where h is Planck's constant ($h = 6.625 \times 10^{-34}$ J/s) and f is the frequency of the EM wave (in hertz). At ultra-high frequencies (UHF) [300 to 3000 MHz], the energies of one photon at the lower band (300 MHz) and at the higher band (3000 MHz) are 1.99×10^{-25} J and 1.99×10^{-24} J, respectively.

In general, kT is the average kinetic

energy possesses by every material particle within the body, where k is the Boltzmann constant and T is the absolute temperature of the particle within the body. The Boltzmann constant in Joul/Kelvin is given by: 1.38×10^{-23}J/K. The absolute temperature of the body (in deg. Kelvin) is given by: 273 + 37 = 310 K. Therefore, kT = $1.38 \times 10^{-23} \times 310 = 4.28 \times 10^{-21}$J. The photon energy ($1.99 \times 10^{-25}$ or 1.99×10^{-24}) in the UHF band is for less than kT (4.28×10^{-21}J) or the bond energy (1.6×10^{-19}J).

These two observations provide strong evidence that there is some form of biological activity in biological materials at nonthermal power levels. The biological effects of EM waves depend upon the electric fields within the tissues. Numerous researchers have reported upon such nonthermal effects of EM radiation on biological materials,[5-13] and some of these results will be reported in the following section.

The biological effects of EM waves depend upon the electric fields within the tissues.

Direct evidence of DNA strand breakage was reported by Sarkar *et al.* in 1994.[19] These researchers investigated the effects upon DNA for exposure to EM field levels deemed safe by the Nonionizing Radiation Committee of International Radiation Protection Association (IRPA). In this study, 20 inbred male Swiss albino mice, approximately five months of age, were exposed to continuous-wave (CW) microwave energy at a frequency of 2.45 GHz for two hours each day in an anechoic chamber. The average power level was 1 mW/cm^2 (for a SAR of 1.18 W/kg). The exposure chamber was tapered, especially designed for frequencies from S to X bands. The anechoic chamber was constructed with pyramidal-shaped radar-absorbing material. The constraining cage was placed symmetri-

cally along the midline of a pyramidal-horn (14.2×10.8-cm) antenna. The distance between the antenna and the cage was 2 m. The antenna was connected to a klystron oscillator through a broadband isolator.

The mice were put into a cage made of 5-mm-thick acrylic, with relative permittivity of 2.2. The cage was covered with a Plexiglas lid having holes for proper ventilation. A total of four control animals and exposed animals (two animals at each interval of 120, 150, and 200 days) were used for DNA isolation from the testis and brain following the method described by Sarkar *et al.*[18] Sarkar and fellow researchers reported significant alteration in the DNA from mice brain and testes for a 7-to-8-kb molecular weight region of the DNA in the hybridization profile and in a densitometric analysis. Gel-track analysis of the control-brain DNA showed a sharp peak (marked 1 in **Figs. 1a and b)** and the appearance of the second peak (marked 2 in the figures) in all of the exposed animals. The hybridization profile for the testes DNA is not a reflection of sharp band differences in this region between exposed and control animals but rather a broadening of the bandwidth. In the exposed animals, there is a change in the peak profile corresponding to 7.7 kb associated with the peak marked 2.

The hybridization profile for the testes DNA is not a reflection of sharp band differences in this region between exposed and control animals but rather a broadening of the bandwidth.

Peak 1 is present in the control and exposed animals, while peak 2 appears in all the exposed animals. From these experiments, it can be concluded that significant alterations in the rats' brain DNA took place due to exposure to the 2.45-GHz CW microwave energy at a level of only 1 mW/cm^2. Lai *et al.*[20,21] reported significant increases in single-and double-strand DNA breaks for rats' brains having been exposed to pulsed microwave energy compared to control subjects. For the experiment, they used a very advanced technique to detect DNA strand breakage which is developed by Singh *et al.*[22,23] at the University of Washington. The technique is known as microgel electrophoresis or the comet assay. Lai *et al.* used spatially-averaged

power densities of 1 and 2 mW/cm^2 to radiate rats in the cage, producing whole-body average SARs of 0.6 and 1.2 W/kg, respectively.

Malyapa *et al.*[24, 25] used an alkaline comet assay developed by Olive *et al.*[26, 27] to measure DNA damage. Upon using a SAR of 0.6 W/kg to radiate rats in the cage, Malyapa *et al.* did not observe DNA strand breakage from microwave exposure. This contrast between their results and those of Sarkar has led to controversy among scientists and engineers.

Further investigation reveals that the comet assay used by Lai *et al.* is more sensitive than the comet assay used by Malyapa *et al.*, due to several factors:

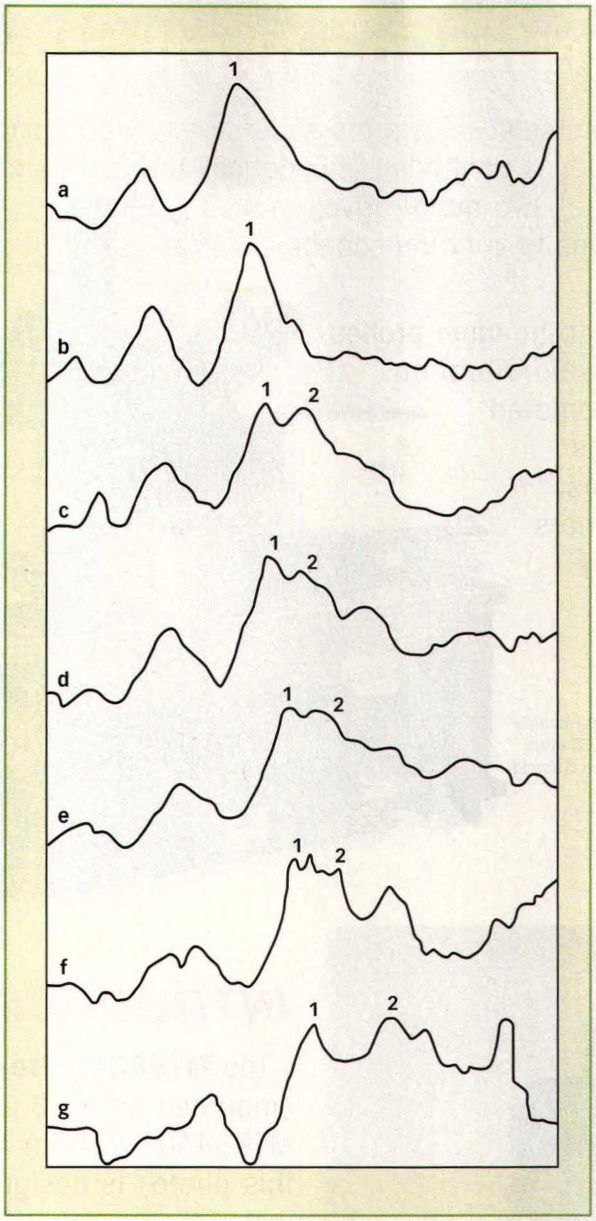
Further investigation reveals that the comet assay used by Lai et al. is more sensitive than the comet assay used by Malyapa et al.

1. complete lysis using highly concentrated salt and two detergents;
2. the use of proteinase Kelvins to remove the positively charges bound protein from the negative-charged DNA strands so that the electrophoresis field produces more migration;
3. the use of antioxidants during electrophoresis;
4. electrophoresis for longer time to allow longer tails to form in the comet (Lai *et al.* have 250 micron tails while Malyapa *et al.* have 40 microtails).
5. use of the YOYO-1 dye by Lai *et al.* to increase sensitivity;
6. the processing approach used to measure the DNA is better in case of Lai *et al.* than in the case of Malyapa *et al.*

Phillips *et al.*[28] reported the results

of comet assays performed to detect DNA single-strand breaks in Molt-4 T-lymphoblastoid cells exposed for short (2- and 3-h) and long (21-h) periods to pulsed signals at the cellular-telephone frequency of 813.5625 MHz. DNA single-strand breaks were measured with alkaline-comet, or single-cell gel electrophoresis assays. Assays were performed using a modification of the technique reported by Singh *et al.*[22] Phillips et al. found significant damage in DNA when integrated-digital-enhanced-networks (iDEN) and time-division-multiple-access (TDMA) signals (2.4 µW/g and 24 µW/g, respectively, for 2 to 21 h) were exposed to the cells.

French *et al.*[29] used a human astrocytoma cell line, U-87MG, with exposure to 835-MHz RF radiation for 20 min., three times per day for seven days, at a power density of 8.1 ±0.8 mW/cm^2. At this power density they observed that the rate of DNA synthesis decreased, and that the cells flattened and spread out in comparison to unexposed culture.

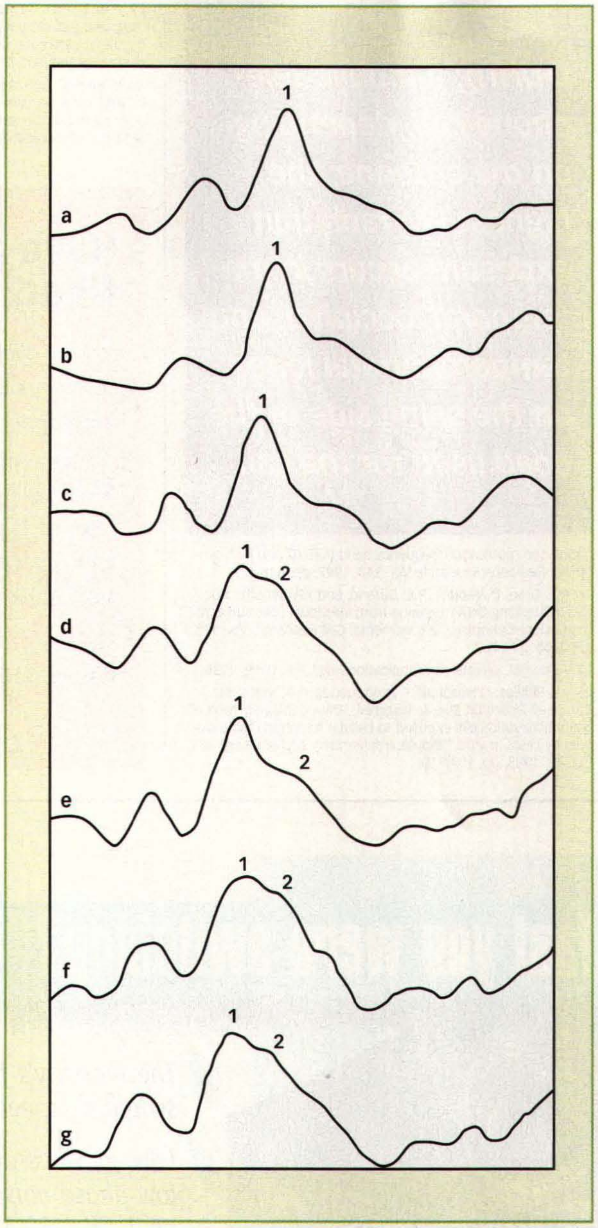


1a. Desitometric analysis of the brian DNA is illustrated here. As shown here, a and b are control DNA, while c to g are DNA from exposed animals. Peak 1 is present in control and exposed animals, while peak 2 is present in all of the exposed animals.

Hardell *et al.*[30] reported the use of cellular telephones as risk factors for brain tumors. During a mobile-telephone call, depending on the antenna, the highest EM radiation exposure occurs in the temporal, occipital, and temporoparietal lobes on the same side of the head used for the call. There is a rapid decline of the radiation dose in the brain, while the other side of the head/brain is only exposed to low levels of EM radiation. Hardell *et al.* reported increased risk of brain tumors in the anatomic area with highest exposure to microwave radiation from a cellular telephone.

In summary, two mechanisms for nonthermal effects have been described. The power levels for these two mechanisms are very low, with no significant

generation of heat occurring. Sarkar and fellow researchers studied the effects for exposure of rat brains to an average power level of 1 mW/cm^2. Although heating was not possible at this power level, it did result in significant alterations in DNA for the rat brains studied. In other studies, Lai *et al.* used power densities of 1 and 2 mW/cm^2 and Malyapa *et al.* used a power density 1 mW/cm^2. In addition, Philips *et al.* exposed Molt-4T cells at a SAR of 2.4 and 24 $\mu W/g$ while French *et al.* used a power density of 8.1 ± 0.8 mW/cm^2 for their experiments. In the lowest-power case, Hardell *et al.* used a power density of less than 1 mW/cm^2. None of these exposure levels produced any temperature effects, although a variety of nonthermal effects were noted due to the presence of the RF and microwave fields. MRF



1b. Desitometric analysis of the testis DNA is illustrated here. As is shown, a to c are control DNA, while d to g are DNA from exposed animals. Peak 1 is present in control and exposed animals, while peak 2 is present in all of the exposed animals.

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***During** a mobile-telephone call, depending on the antenna, the highest EM radiation exposure occurs in the temporal, occipital, and temporoparietal lobes on the same side of the head used for the call.*

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UMTS Networks Architecture, Modility and Services

HEIKKI KAARANEN, ARI AHTIAINEN, LAURI LAITINEN, SIAMAK NAGHIAN, AND VALTTERI NIEMI

THE WORLD'S first public GSM call was made on July 1, 1991 in a city park of Helsinki, Finland. That event is now regarded as the birthday of the 2G mobile telephony. GSM has been an overwhelming success, which was difficult to predict at that early stage. In the past 10 years, GSM has truly become a global system for mobile communications. There are now cellular-phone penetration rates exceeding 50 percent in many countries and approaching 80 percent in the Nordic countries. In 10 years, GSM has brought us onto the footstep of the 3G mobile communications system—UMTS.

The latest estimates show that packet-switching traffic in the mobile core networks will exceed circuit switching by 2005. This transition is enabled by the UMTS system, which makes it possible for the network operators to provide equally strong circuit-switched and packet-switched domains to meet the speed and capacity demands. Most voice and time- critical data services may still use circuit switching, while less time-sensitive data passes through the UMTS mobile packet-core network.

UMTS Networks Architecture, Modility and Services is about UMTS networks as a 3G platform for mobility and services. It aims to provide a comprehensive overview of the system architecture and its evolution and to serve as a guidebook to those who need to study the specifications from 3GPP. The content of the book is divided into three parts. The first part consists of Chapters 1 and 2, which serve as an executive summary of the UNITS system.

Chapter 1 introduces the UNITS technical and service architecture and key system concepts. An introduction to 3G network architecture, a review of a conceptual network model, and structural network architecture are included. Also reviewed are resource management architecture and UMTS service and Bearer Architecture. Chapter 2 is an illustrated story about mobile network evolution from second-generation GSM into the first UNITS release and beyond that toward full IP mobility networks. The chapter also includes a review of network and service evolution.

The second part consists of Chapters 3 through 8, which examine the radio-access and core networks in more detail, explaining the functions and services provided to the end users. Chapter 3 concentrates on UNITS radio communications and provides the fundamentals of cellular radio, which are necessary for understanding the WCDMA radio-access system as part of the UNITS network. Chapters 4 and 5 present the functional-split and system-management aspects distributed among the UNITS network elements in the radio-access and core-network parts. Topics in Chapter 4 include UTRAN architecture, modulation methods, cell capacity, and network controllers. Chapter 5 reviews UMTS core networks, core-network architecture in 3GPP, and mobility management.

Chapter 6 provides an overview on the UNITS user equipment and focuses on those aspects which are most visible to the rest of the UNITS network. In Chapter 7, the UNITS network is examined as a network for services. Topics include UNITS service capabilities including WAP and CAMEL. Location-based services are introduced as a value-added service platform and application examples with different QoS characteristics are discussed. The advanced security solutions of the UNITS network are discussed in Chapter 8.

Chapters 9 and 10 form the third part of the book. In these chapters, a protocol-oriented view is taken, describing the system-wide networking between the different architectural elements. Chapter 9 first elaborates the basic UNITS protocol architecture and then introduces the individual system protocols one by one. Chapter 10 returns to the network-wide view by showing selected examples of system procedures, which describe how the transactions are carried out across the UMTS network interfaces under the co-ordination of the system protocols. (2001, 302 pp., hardcover, ISBN: 0471-48654-X, $85.00) **John Wiley and Sons, Ltd., 605 Third Ave., New York, NY 10158-0012; (212) 850-6000, FAX: (212) 850-6088, Internet: www.wiley.com.**

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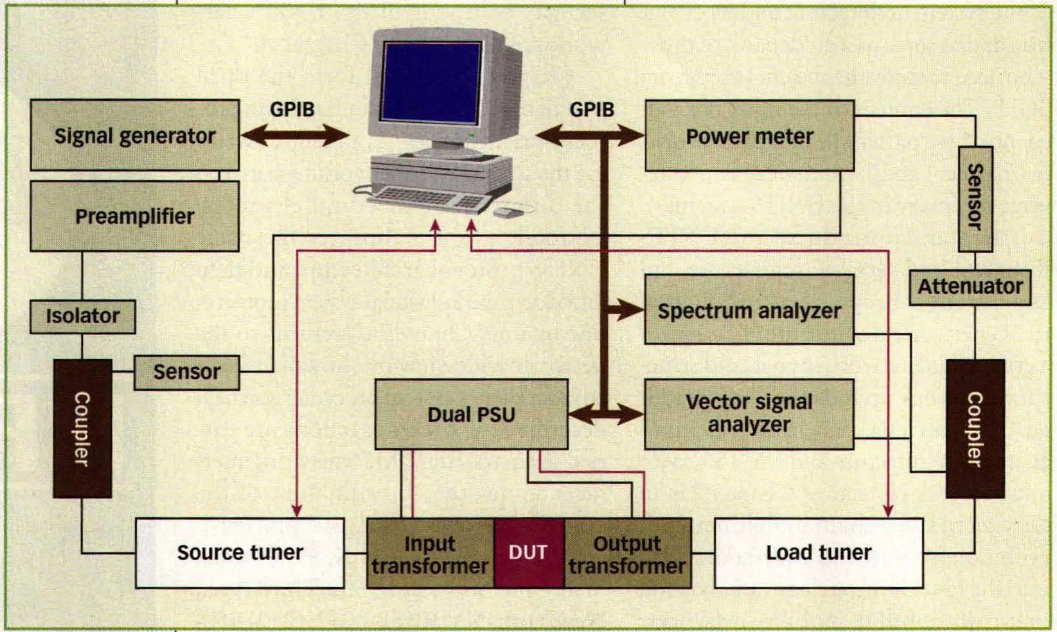

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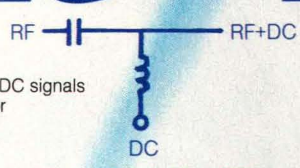

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been developed to maximize the amount of information carried by relatively narrowband signals. Examples of these formats include wideband code-division multiple access (WCDMA) and orthogonal frequency division multiplexing (OFDM). Both impose severe linearity constraints on the design of a communications power amplifier (PA). Traditional tuning techniques using stub tuners, tuned circuitry, and device de-embedding have been successfully used for many years. However, these techniques do not provide a complete picture of the trade-offs between parameters such as efficiency and linearity, or

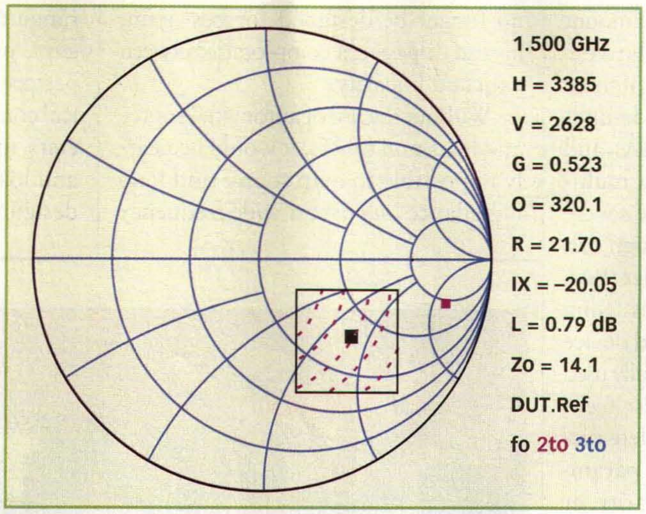
Designers generally assume linearity to be a function of the load impedance presented to the amplifier.

of the sensitivity of the parameters to impedance variations. Designers generally assume linearity to be a function of the load impedance presented to the amplifier. But significant improvements are possible when the source and load impedances are simultaneously optimized for linearity.

In part, this is due to the lack of forward and reverse isolation of high-power, high-frequency devices where changes in the output match reflect back to the input. A common design approach for optimum gain is to optimize the load-impedance linearity, then optimize the source impedance. Impedances applied to an RF power transistor can be chosen for maximum load power, maximum gain, or maximum linearity, but rarely can all three parameters be maximized simultaneously. Output linearity performance can be significantly affected by the choice of impedance presented to the base or gate of an active device. However, this complicates the design process, since the input-matching network can no longer be designed for best gain, instead delivering a compromise between gain and linearity.

Without the use of automated source and load-pull tuners, it would be nearly impossible to map source and load impedance loci over a wide frequency range to achieve the best balance of gain, power, efficiency, and linearity performance. The generation of graphical contours for various system parameters, measured as a function of source and load impedances, is invaluable in designing reliable RF PAs targeted for

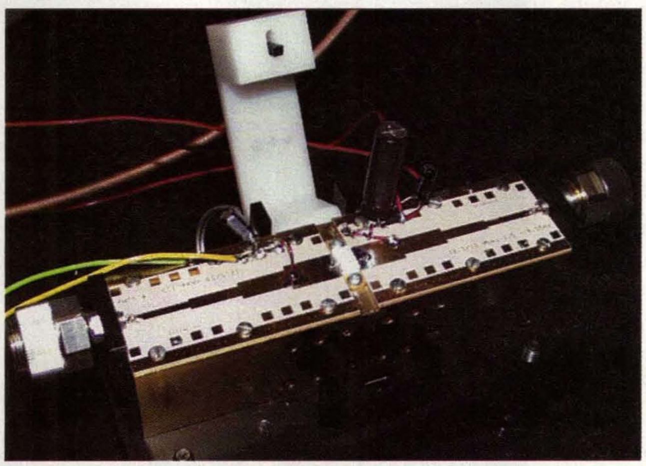
specific performance criteria. Without this type of information, the effects of production-line variations, as well as variations in impedance, may be completely unknown. Without this information, there is a strong temptation to over-design an amplifier, for example through use of unnecessarily complex linearization techniques, to ensure that it meets its required performance levels.



2. Parameter contours can be saved to digital files by selecting areas of interest from input and output planes.

Improvements in adjacent-channel power leakage (ACPL) on the order of 6 dB can be achieved through optimum selection of source and load impedances without recourse to linearity correction. The loss of efficiency by seeking such improvement through power reduction would be dramatic. For high PA design, the cost savings in linearity correction, power-supply design, and heat removal are significant. The use of automated load-pull equipment based around computer-controlled programmable tuners enables the generation of parameter contours under real high-power conditions and supports fair comparison of devices. The tuners can directly apply a wide range of impedances to the input and output of a device. If the expected device impedances are outside the regions to which the tuners can directly tune, transmission-line impedance transformers on the device-under-test (DUT) fixture can be used to center the tuners to the required range.

A source and load-pull measurement system is initially calibrated so that the impedances presented to a DUT at the reference planes of the text fixture are known for all tuner settings. During a source or load-pull run, the tuners are automatically controlled to visit a grid of impedance points within a user-definable region of the impedance plane. For each impedance setting, a user-definable set of device-system performance data is recorded using the outputs from network analyzers, power meters, vector-signal analyzers, and other associated test equipment. External measurement equipment is remotely controlled through general-purpose-interface-bus (GPIB) connections. **Figure 1** shows a typical load-pull setup, configured in this case to measure the adjacent-channel leakage ratio (ACLR) of a device with CDMA modulation.



3. This test fixture employs a pair of stepped printed-circuit impedance transformers.

Calibration Procedure

As part of the calibration procedure, two-port scattering (S)-parameter files are generated for the tuners and all other passive circuitry and these are stored as a "set-up" file. This calibration procedure enables all parametric data to be referred to reference planes close to the DUT package boundary. The equipment user interface features two Smith charts representing the impedances applied to the input and output planes. In high-resolution mode, the measurement comprises approximately 800 calibrated impedance points distributed over the complete Smith chart. The input and output tuners are typically able to directly generate impedances corresponding to reflection coefficients up to approximately 0.96. Areas of particular measurement interest can be sectioned out from the input or output planes, stored as a file, and used to generate parameter contours **(Fig. 2)**.

A three-part fixture is used for the measurement of power transistors having in-line transmission-line tabs consisting of a center section where the device is mounted and two half-sections for mounting printed-circuit boards (PCBs) containing printed transmission lines. Depending on the expected device impedances, the transmission lines may or may not be used to perform impedance transformation. To calibrate the system in either case, a through-reflect-line (TRL) calibration is performed. To carry out a TRL calibration, the center section of the test fixture is removed, and a calibration is performed with the two halves of the test fixture cascaded together.

A small section of transmission line of known impedance is then inserted between the two halves in place of the DUT and another reflection calibration is performed. Finally, the two halves of the test fixture are separated and a reflect calibration is performed on each half. The attenuation of the transmission-line sections is not critical, but the optimum electrical length of the line section

should be approximately 90 deg. at the center of the test-frequency range. Adequate measurement accuracy is assured if the electrical length is within 30 to 150 deg. at any frequency in the range of interest. The characteristic impedance of the line establishes the measurement reference impedance. The impedance of the fabricated transmission lines can be calculated, but may be subject to minor variation due to substrate height, dielectric constant, and chemical-etching tolerances. Ultimate accuracy is assured using a time-domain-reflectometer (TDR) measurement to validate the calculated impedance.

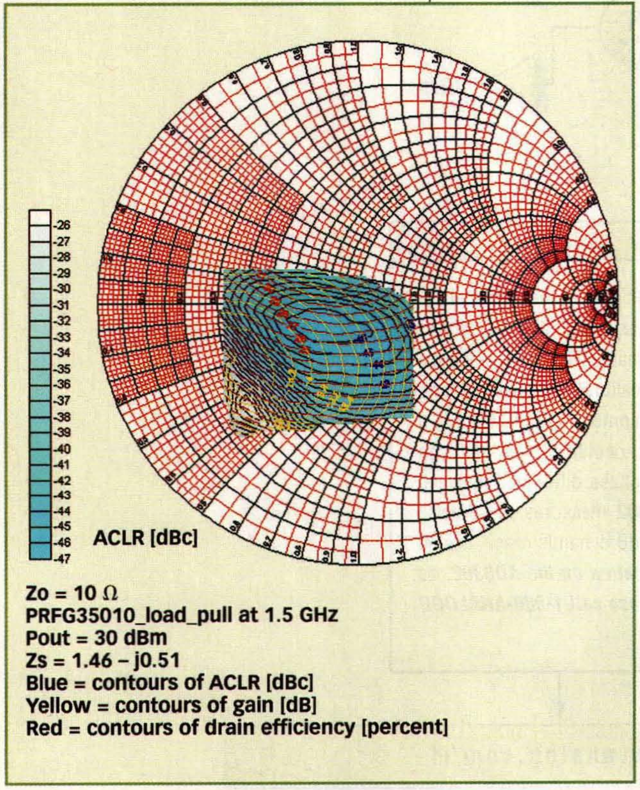
This calibration procedure establishes measurement-reference planes at the device-package end of each half-section. The procedure uses a twelve-term error-correction model to remove the systematic errors, such as insertion loss and connector discontinuities associated with the fixture. This results in a pair of two-port S-parameter files saved as setup files. Similar setup files are created for all other signal-path elements used in the measurements. This calibration procedure enables effective de-embedding of the device for applied power and impedances. **Figure 3** shows an example of a test fixture employing a pair of stepped printed-circuit impedance transformers. This fixture can be used characterize a device down to levels of approximately 3 to 5 Ω over a one-and-one-half-octave frequency range.

After calibration, a load-pull search with coarsely spaced impedance points is typically performed to determine where the optimum impedance boundaries for gain, linearity, and efficiency are located. The load-tuner impedance is approximately set to an optimum value and a coarse source-pull search carried out. The procedure is repeated for the source tuner and another load-pull search is performed. In this manner, it is possible to achieve simultaneous optimization of source and load impedances within several source and load-pull iterations.

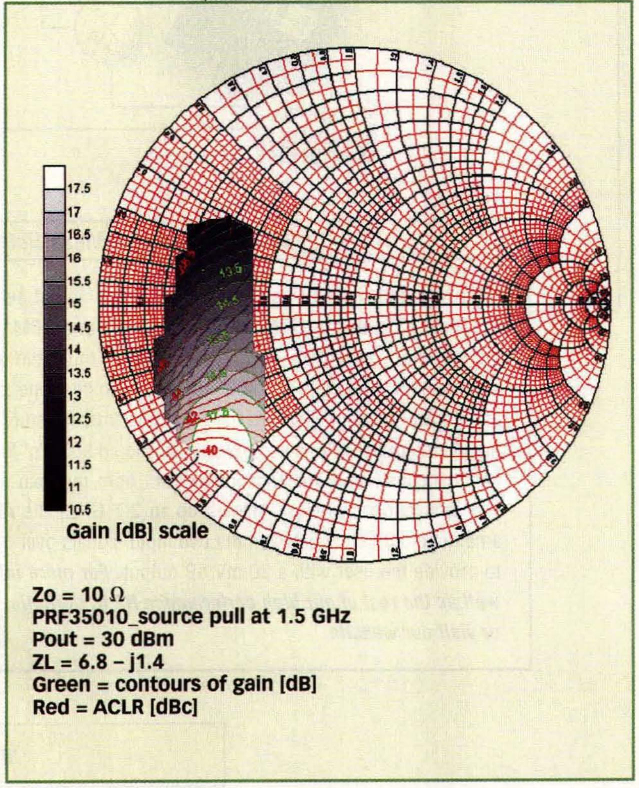
Once approximate source- and load-impedance points are established, subsequent source and load-pull searches with finely spaced impedances steps are carried out to examine the regions of interest in more detail. Contour results such as gain, 1-dB compression, efficiency, and adjacent-channel power ratio (ACPR) may be plotted on a Smith chart. An automated power sweep is used to increase the input power in predetermined level intervals to determine the 1-dB compression point. **Figure 4** is an example of a load plane plot and clearly shows that the impedance points of optimum gain, efficiency, linearity, and ACLR are far from coincident.

Test Data

Figure 5 shows a series of source-pull gain contours for a model PRF35010 pseudomorphic high-electron-mobility transistor (pHEMT) at 1.5 GHz. The device was terminated with constant load impedance of 6.8 – j1.4 Ω, driven with a WCDMA modulated signal, and maintained at a constant output power of +30 dBm. The chart is normalized to a characteristic impedance value of 10 Ω. The lighter shaded area of contours clearly shows a region of maximum gain of



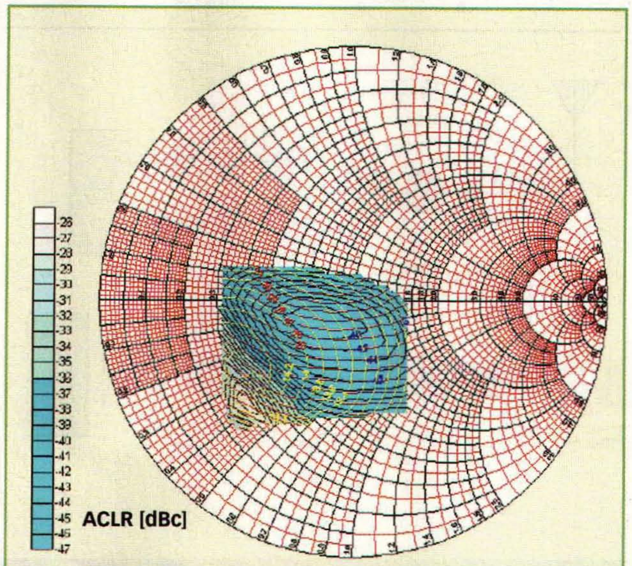
4. This is an example of a Smith chart showing load impedances for optimum gain, efficiency, and ACLR performance.



5. These gain contours were derived from source-pull measurements on a commercial pHEMT device at 1.5 GHz.

17.5 dB, which would be achieved by conjugate matching with an impedance on the order of 1.5 – j4.0 Ω. This is a generally adopted solution resulting from a practical search using stub tuners or tunable and demountable circuitry as a means of assessing performance against certain applied impedances.

The ACLR contours in Fig. 5 show that ACLR of approximately –40 dBc is achievable if the device input is matched for maximum gain. In addition, an unexpected result observable from this Smith chart is that a 6-dB improvement in ACLR is possible if 3-dB gain is sacrificed by mismatching the input. To achieve such improvement from an amplifier conjugate matched at its input would require significant additional power reduction from the 1-dB compression point, resulting in significant loss of drain efficiency.



6. These load-plane ACPR, gain, and efficiency contours result for the commercial pHEMT after input-match optimization.

Once the source impedance is optimized for improved ACLR, the variation in ACPR, gain, and drain efficiency with load impedance for constant output power can be mapped out. The result is the contour map shown in **Fig. 6.** The central (blue) dense area indicates the boundary of improved ACLR of –46 dBc, intersected by (yellow) contours of gain (13.4 to 13.8 dB) and (red) drain efficiency of 27 to 28 percent. This performance is centered on a drain-load impedance of 6.8 – j1.4 Ω.

The use of automated source and load-pull techniques enable significant improvements in realizable RF power-transistor linearity and efficiency compared with traditional design techniques. The technique also reveals critical information on design sensitivity and allows stable, reliable, high-yield designs to be realized. Although the examples in this article are from the telecommunications sector, the techniques described are equally applicable to PA designs for a wide variety of RF markets, including satellite, broadcasting, avionics, and military applications.

Recently, prematching tuners have become available, which lessen the need for restricted bandwidth on-fixture impedance transformations and enable accurate measurements of devices down to impedances less than 1 Ω. Further improvements to amplifier linearity and power-added efficiency (PAE) can occasionally be achieved by reflecting harmonic energy back into the output stage. This requires independent control of the phase of unity gamma reflections at second and third harmonics. Harmonic tuners, which generate reflection coefficients in the range 0.95 to 0.99 at second- and third-harmonic frequencies, are also available to address these design requirements. MRF

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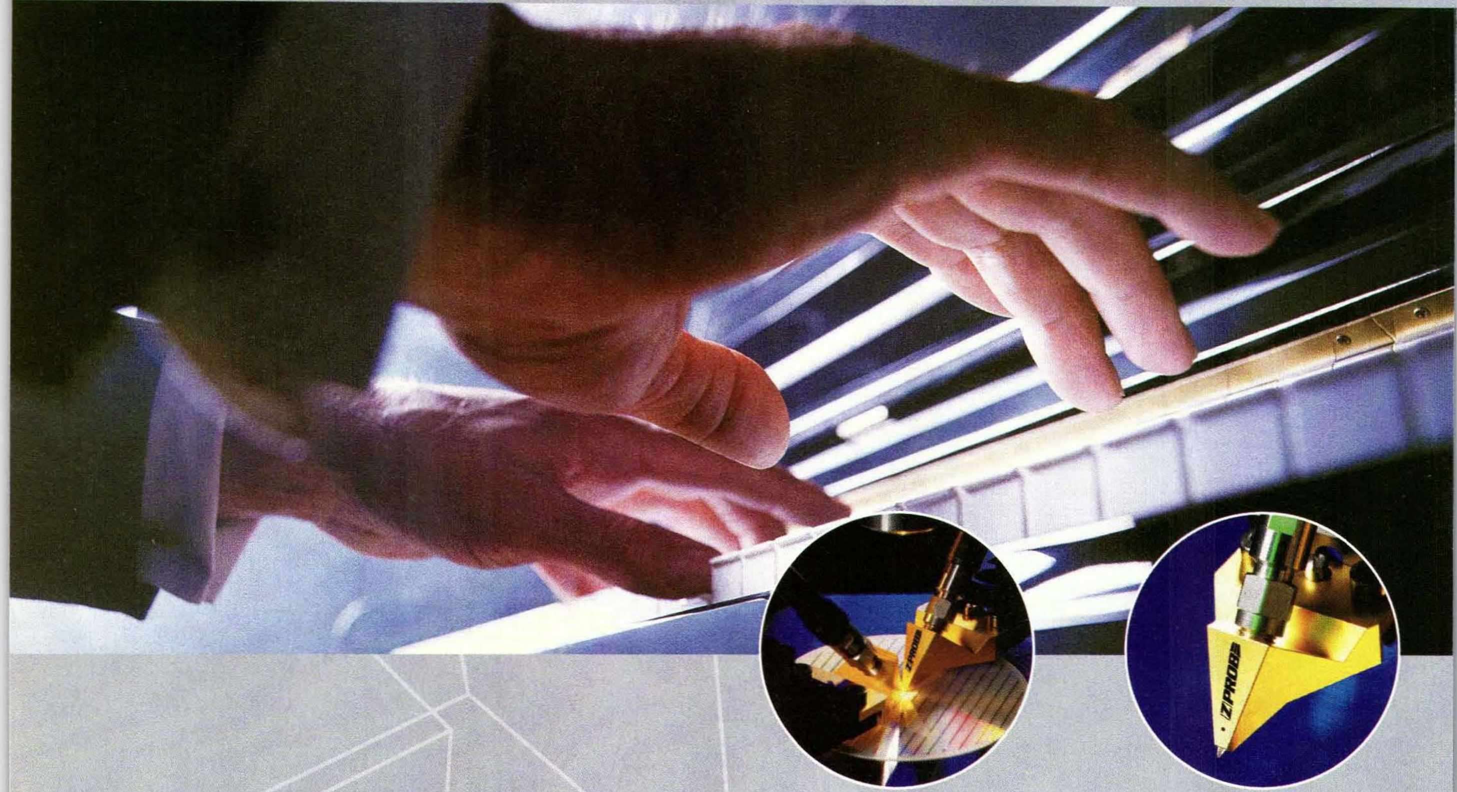

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DWDM Fundamentals, Components, and Applications

JEAN-PIERRE LAUDE

PRESENT AND EXPECTED future consumer data demands continue to push the need for increased bandwidth. In optical systems, the use of different wavelengths for multiple signals in a common fiber is essentially the basis for DWDM. Of course, the optical-communications technique is somewhat more complicated than that simple explanation, and those wishing to know more are advised to refer to *DWDM Fundamentals, Components, and Applications* by Jean-Pierre Laude. The author holds more than 50 patents, an engineering degree in optics from the Ecole Superieure d'Optique de Paris, and a Ph.D. in spectroscopy from the University of Paris at Orsay.

The volume's more than 200 pages offer comprehensive coverage of the principles, technologies, standards, and applications for DWDM. The author has organized the text in only eight chapters, the first of which is a two-page introduction to DWDM. Subsequent chapters review demultiplexers; optical sources and wavelength converters; WDM techniques and optical amplification; routers, cross-connect circuits, and add-drop methods; the limits of the technology imposed by optical nonlinearities in optical fibers; and how to apply DWDM technology to modern telecommunications networks.

The book is generously illustrated with basic graphics and plots of performance, and written simply enough to satisfy the educational needs not only of technical professionals, but of business managers as well. *DWDM Fundamentals, Components, and Applications* clearly explains how DWDM components, devices, and networks operate and provides a good first look into the configuration and design trade-offs of current DWDM components and systems. Even readers with limited knowledge of engineering and optics should come away with a better understanding of DWDM components and systems after reading this book and, for the more "seasoned" reader, the text is ripe with references for sources of additional and more-complex information on DWDM. (2002, 282 pp., hardcover, ISBN: 1-58053-177-6, $85.00.) **Artech House, Inc., 685 Canton St., Norwood, MA 02062; (781) 769-9750 ext. 4030, FAX: (781) 769-6334, e-mail: artech@artechhouse.com, Internet: www.artechhouse.com.**

Avoid Frequency Extrapolation Errors

Relying on frequency extrapolations of known measured data can lead to erroneous simulations of S-parameter responses in commercial simulators, for active and passive models.

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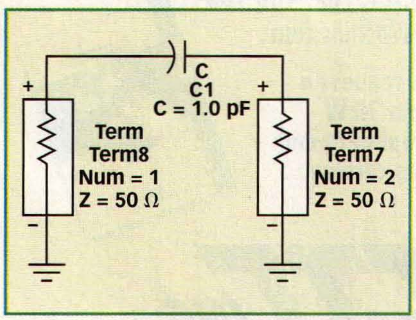
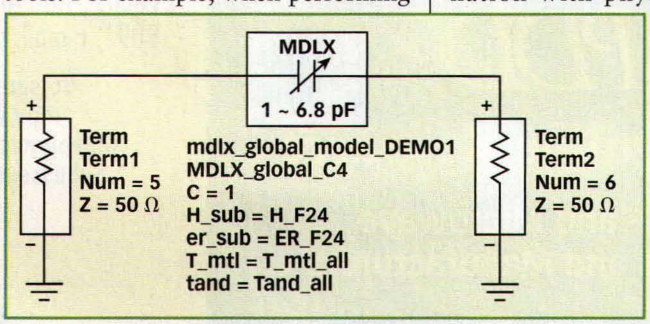
Extrapolation of unknown frequency data for computer-aided-engineering (CAE) models can translate to misleading results. Measured scattering (S)-parameter data sets for passive components and transistors are fine for linear simulations that do not go outside the frequency limits of the measurement. But nonlinear harmonic-balance simulations, by their nature, require simulations at 5 to 10 harmonics of the desired operating frequency. Commercial simulators can extrapolate these higher harmonic information based on known lower-frequency measured data for active and passive devices, but these extrapolations often lead to nonphysical and inaccurate simulation results. Examples for this effect will be presented for a surface-mount capacitor and a high-electron-mobility transistor (HEMT).

Broadband frequency sweeps are often necessary when designing RF/microwave circuitry using CAE tools. For example, when performing nonlinear circuit simulations such as harmonic-balance simulations, it is common to include 5 to 10 harmonics of the fundamental design frequency. Components contained within a circuit schematic are also required to be valid at DC if these components are part of a DC path (e.g., bias paths for transistors). An ideal situation should include circuit models for all of the circuit components that are at best accurate, and at the least, well-behaved from DC through five or ten times the anticipated frequency of interest.

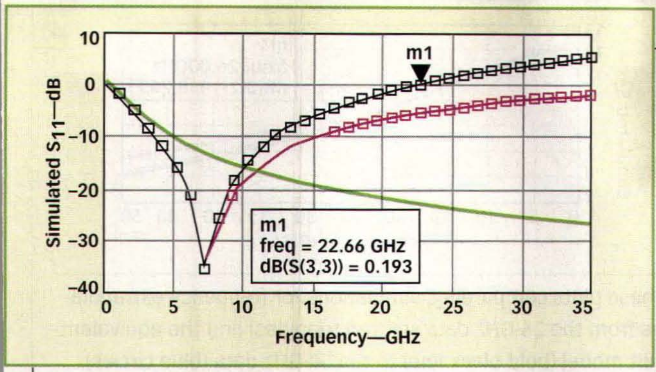
Properly performed S-parameter measurements can be used in combination with physically motivated equivalent-circuit topologies to produce valid broadband circuit models. In some cases, the S-parameter measurements themselves are a useful substitute for a physical model. The use of these measure-

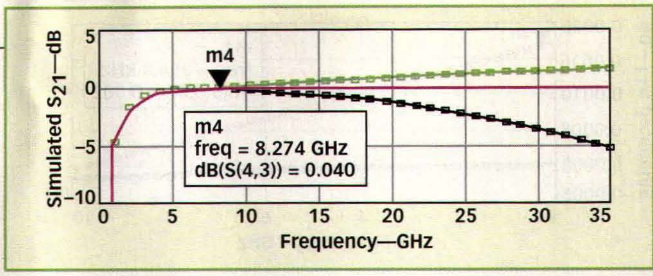


1. This schematic diagram shows an ideal 1-pF capacitor.



2. This substrate-scalable global capacitor equivalent-circuit model has been set to 1 pF.





4. These simulated S_{21} results are shown for an ideal capacitor (thick black line), equivalent-circuit model (magenta circles), and as extrapolated from a data set truncated at 6 GHz (blue squares).

3. These simulated S_{11} responses for an ideal capacitor (thick black line), equivalent circuit model (magenta circles), and as extrapolated from a data set truncated at 6 GHz (blue squares).

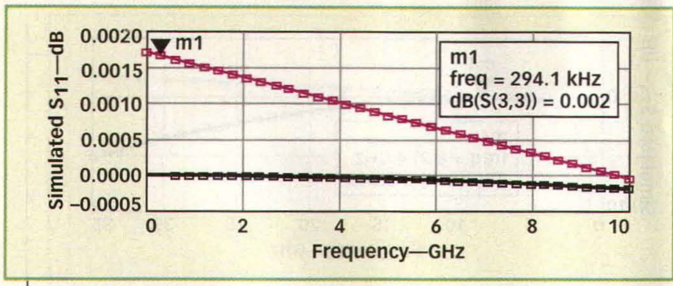
ments avoids errors associated with model extraction and/or curve/data fitting. In a similar way, electromagnetic (EM) simulation can be used to generate S-parameter data sets for arbitrary two- and three-dimensional (2D and 3D) passive structures.

Some simple examples may be useful in demonstrating the potential pitfalls that can occur during broadband and/or DC simulations. The particular situation addressed is that when S-parameter-based "data-set models" are used, in comparison to equivalent circuit models, in these simulations.

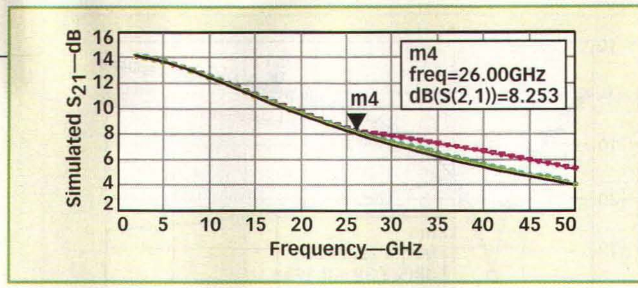
The first modeling example features the effects of performing passive simulations using different sets of data and a typical high-frequency component model, a 1-pF capacitor. Simulations will be compared for:

1. An ideal 1-pF capacitor in a series two-port configuration **(Fig. 1)**.
2. A measured S-parameter data set for a 1-pF capacitor, with data from 0.05 to 6.00 GHz. The original data set was extended through 10 GHz but truncated for the purpose of this example.
3. A substrate-scalable equivalent-circuit model for a 1-pF capacitor extracted from the measured (through 10 GHz) S-parameter data set prior to truncation **(Fig. 2)**.

The simulated S_{11} responses for the three modeling approaches are shown in **Fig. 3**. The frequency is swept from 0.05 to 35 GHz, which would provide coverage through the sixth or seventh harmonic of a 5-GHz wireless-local-area-network (WLAN) design. In this example, the response



5. These simulated S_{11} results are shown for an ideal capacitor (no marker), equivalent-circuit model (magenta circles), and as extrapolated from a data set truncated at 6 GHz (blue squares).



6. These plots compare S_{21} simulations for frequency extrapolations from the 26-GHz data set (red triangles) and the equivalent-circuit model (bold black line) to the 50-GHz data (blue circles).

of the ideal capacitor is inaccurate after approximately 3 GHz, while the equivalent-circuit model tracks the measurement data through 7 GHz. Most significantly, the extrapolation of the measured data set beyond its upper limit (6 GHz) leads to a *non-physical* result of $S_{11} > 0$ dB past 22 GHz. The response of the equivalent-circuit model, on the other hand, has a well-behaved monotonic increase toward (but not exceeding) 0 dB.

The simulated S_{21} responses for the same 0.05 to 35 GHz sweep are shown in **Fig. 4.** In this case the extrapolated S_{21} from the measurement data set goes above 0 dB (indicating gain for a passive) at 8 GHz, which is only slightly beyond the upper limit of the S-parameter file.

The low-frequency S_{11} responses of the schematics are shown in **Fig. 5.** As would be expected, the ideal capacitor response is indistinguishable from the equivalent-circuit model prediction since any parasitic effects are negligible. However, the extrapolation of the measurement data set once again leads to a non-physical result of $S_{11} > 0$. This result can lead to convergence errors during harmonic-balance simulations (e.g., analysis of microwave mixers and frequency converters). Similar problems arise in inductor, filter, and passive matching network extrapolations.

To understand the active-device simulation issue, HEMT S-parameters generated within a commercial CAE simulator were compared for:

1) A measured data set whose upper frequency limit is 26 GHz.

2) A measured data set whose upper frequency range is 50 GHz.

3) An accurate broadband equivalent-circuit model for the HEMT.

Figure 6 offers a comparison of S_{21}

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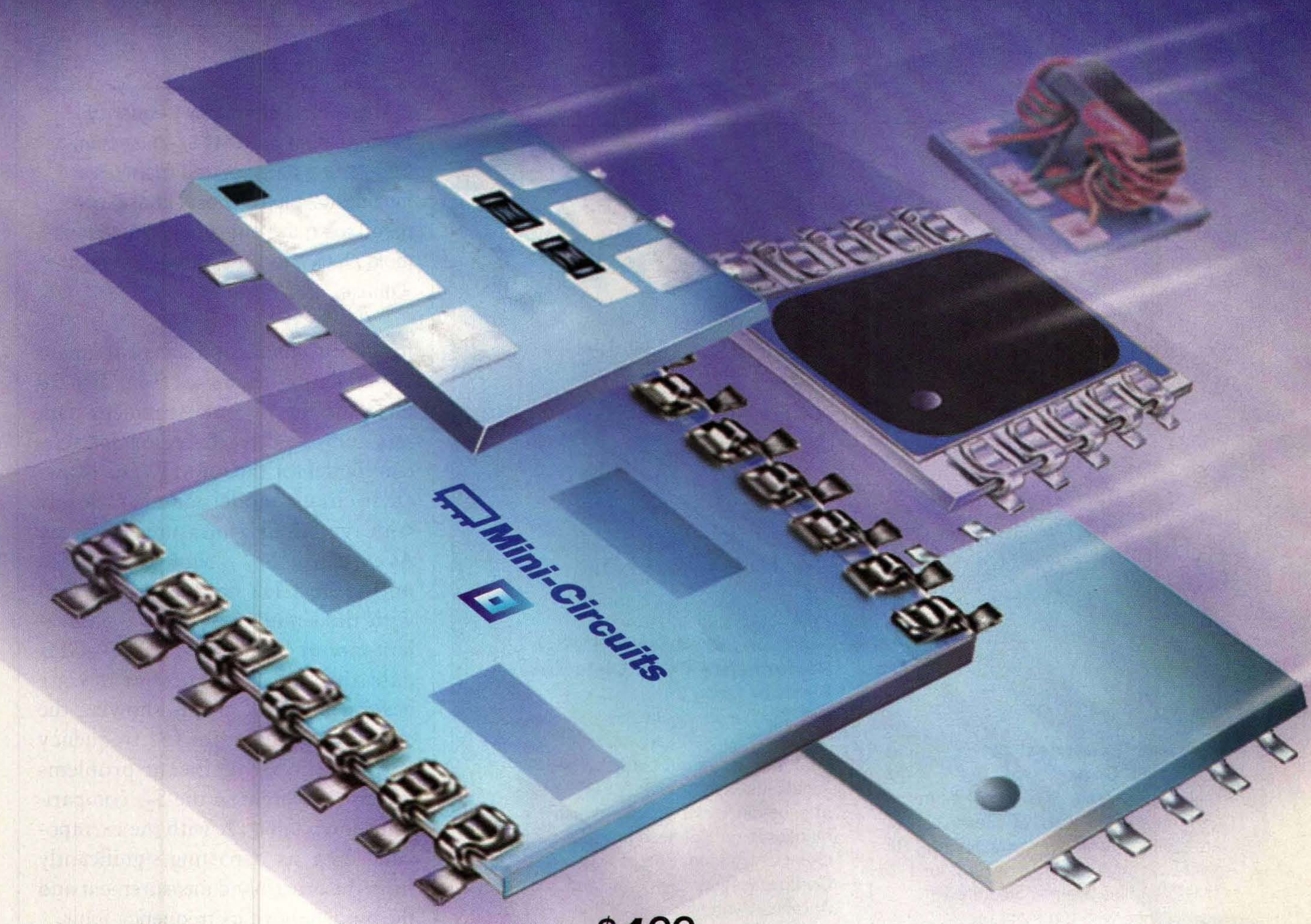

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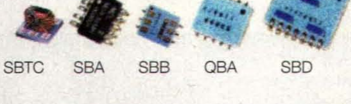

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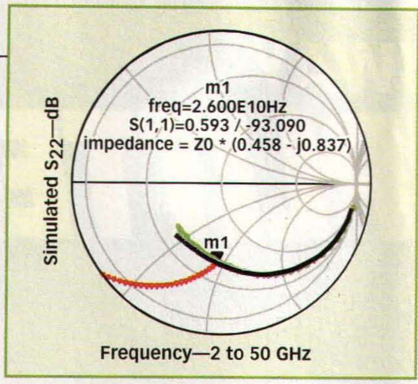
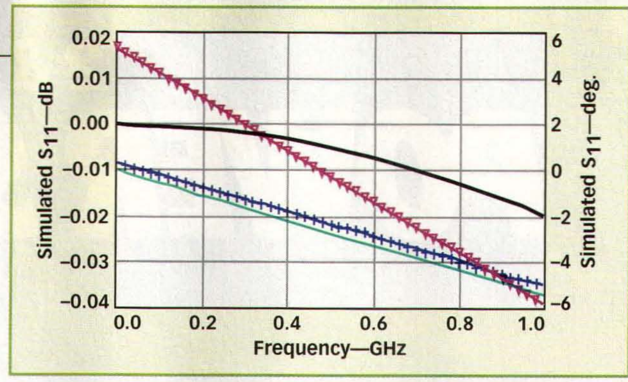


7. These plots show a HEMT transistor S_{22} comparison of frequency extrapolation from 26-GHz data set (red triangles) and circuit model (bold black line) to 50-GHz data (blue circles).



8. These plots show low-frequency extrapolation of HEMT S_{11} data from a 2-to-26-GHz data set (red triangles = magnitude, blue circles = phase) compared to equivalent-circuit simulations (bold black line = magnitude, green line = phase).

forward-transmission swept-frequency simulation responses from 2 to 50 GHz for the two measurement data sets compared to the equivalent-circuit model for the HEMT. The equivalent-circuit model matches closely with the 50-GHz measured data set. As expected, the S_{21} simulations based on the 26-GHz data set agree well with the simulations for the equivalent-circuit model and the 50-GHz data through 26 GHz, but deviate beyond 26 GHz, clearly showing the errors resulting from the frequency extrapolation. Significant problems are also apparent in the S_{22} comparison shown in **Fig. 7**, with the extrapolated data set departing significantly from the broadband measurement and the model above its frequency limit.

Since low-frequency data can be essential to the biasing of a HEMT device, low-frequency extrapolations were made for S_{11} **(Fig. 8)** simulations of the HEMT. The basic 2-to-26-GHz data set was extrapolated to the DC-to-1-GHz range and compared to the results for the equivalent-circuit model. Problems with the data extrapolation are evident in the S_{11} data near DC. For the extrapolations, the magnitude of S_{11} rises above 0 dB, indicating an instability not shown in the model simulations. The phase of S_{22} (not shown) also exhibits a nonphysical result, with positive reactance below 200 MHz. MRF

EDITOR'S NOTE

Thomas Weller, Ph.D. and Lawrence Dunleavy, Ph.D. are also affiliated with the Dept. of Electrical Engineering at the University of South Florida in Tampa, FL.

Fast and Efficient Algorithms in Computational Electromagnetics

WENG CHO CHEW, JIAN-MING JIN, ERIC MICHIELSSEN, AND JIMING SONG

Fast and Efficient Algorithms in Computational Electromagnetics documents recent advances in computational EMs in the manner of a monograph. For those who intend to perform research in this area, this book will be an excellent starting point, as it focuses on linear problems associated with Maxwell's equations.

Chapter 1 introduces EM analysis and explains how the field has evolved into computational EMs in the last few decades. It also introduces, in a very simplified manner, the recent fast algorithms developed to solve Maxwell's equations.

FMM and MLFMA in 2D are introduced in Chapter 2. Interpolation, truncation, and integration efforts are discussed. An attempt is also made to relate FMM to group theory, and to the inherent symmetry of space. Chapter 3 describes the 3D version of FMM and MLFMA and demonstrates the application of the fast algorithm to real-world problems.

Distributed-memory parallelization of MLFMA, encapsulated in a code known as ScaleME is reviewed in Chapter 4. The parallelization of MLFMA on a distributed memory machine is not an easy task, because different parts of the computation may reside on different processors.

Low-frequency solution of Maxwell's equations using fast algorithms is presented in Chapter 5. This chapter describes the treatment needed for FMM and MLFMA to prevent their catastrophic breakdown at low frequencies. It also describes a method to apply the LF-MLFMA based on RWG, wire, and wire-surface bases while the intrinsic expansion bases are still the loop-tree-star bases.

Different error issues involved when solving surface integral equations related to Maxwell's theory are reviewed in Chapter 6. Discretization error due to the use of basis functions and integration error by replacing integrals with summation are included. The chapter also discusses deconditioning due to the near-resonance problem and the low-frequency breakdown problem.

Chapter 7 deals with a recent topic of intense interest in differential equation solvers—the theory of PML. The concept of complex coordinate stretching is discussed. PML is generalized to curvilinear coordinates, as well as to complex media. In this chapter, stability issues related to PML are studied and a unified analysis of various PML formulations using differential forms is included. Chapter 8 addresses the issue of efficiently solving the forward and inverse problems for buried objects using FFT-based methods. The detection of buried objects usually involves loop antennas, and the forward problem involving the solution of loop antennas over a buried object is discussed in detail. Moreover, recent advances in different inversion algorithms are also described.

Solving the penetrable problem at very low frequencies is the focus of Chapter 9. The low-frequency problem encountered in Chapter 5 for metallic objects also occurs for dielectric and lossy material objects. This chapter describes a way to solve this problem so that the solution of integral equations remains stable from zero frequency to microwave frequencies.

Chapter 10 describes an algorithm to solve 3D waveguide structures using numerical-mode matching, but using the finite difference method. The spectral Lanczos decomposition method is used to find the modes. An algorithm with O*(N)* memory complexity and $O(^{NI-5})$ computational complexity is achieved. Chapter 11 addresses the problem of solving the volume integral equation concurrently with the surface integral equation. This is particularly important when dealing with structures having metals as well as dielectric materials.

Chapter 12 deals with solving axially symmetric, BOR geometry using FEM. This reduces a 3D problem to 2D, greatly enhancing the efficiency of the solution. Material-coated and metallic objects are considered. Hybridization in computational EMs is the topic of Chapter 13. Hybridization between FEM and ABC is discussed along with BIE, MLFMA, AABC, and SBR. Hybridization between MOM and SBR is also considered. AABC is a promising method of hybridizing FEM with fast solvers for the future.

Chapter 14 presents different higher-order methods for computational EMs, surface integral equations, and FEM. In addition, the efficient coupling of higher-order methods to fast solvers such as MLFMA is discussed. Chapter 15 discusses AVVE for broadband calculation in EMs. Illustrations of this acceleration technique for broadband calculation are provided for metallic antennas, wire antennas, dielectric scatterers, and microstrip antennas.

Microstrip structure analysis on top of a layered medium is detailed in Chapter 16. The derivation of the layered-medium Green's function together with its numerical approximation by the complex images is discussed. The use of the fast-frequency sweep method, adaptive integral method, and MLFMA to accelerate solution speed is studied. A higher-order method to improve solution accuracy is also demonstrated.

Chapter 17 reviews SDFMM to accelerate the solution speed of quasi-planar structures. For this class of structures, this method reduces the computational and memory complexity of MLFMA from O*(N log N)* to O*(N)*. Applications to scattering from random rough surfaces, quantum-well gratings, and microstrip antennas are demonstrated with this analysis method.

Chapter 18 elaborates on the PWTD algorithm, which is an ingenious way of arriving at the time-domain equivalent of FMM and MLFMA. The integral equation is solved using the MOT method. Stability and accuracy issues are carefully analyzed in this chapter. The two-level and multilevel algorithms are presented and demonstrated with examples.

Chapter 19 further develops PWTD for large-scale and real-world applications. The use of PWTD with MFIE, EFIE, and CFIE is illustrated. (2001, 931 pp., hardcover, ISBN: 1-58503-152-0, $129.00.) **Artech House, 685 Canton St., Norwood, MA 02062; (781) 769-9750, FAX: (781) 769-6334, Internet: www.artech-house.com.**

Ensure accuracy in time-jitter measurements

THE ANALYSIS OF jitter is structured on repetitive measurement of the time between two signal edges. One does not have to make absolute amplitude measurements and it is therefore possible to bandwidth limit the input signal. If the edges have equal risetime, this limitation of the frequency can increase risetime without causing instability in the relative time of the signal edges. Filtering ensures that the edge is effectively sampled without cyclic time-measurement degradation.

A nine-page application note entitled "Accuracy In Time Jitter Measurements with LeCroy Oscilloscopes" from Lecroy Corp. (Chestnut Ridge, NY) attempts to provide an understanding of signal digitizing and measurement methods that contribute to the accurate characterization of jitter errors. The article provides an explanation of the five stages in the measurement chain that are required to ensure optimum measurement accuracy, as well as discussing the importance of sample rate in the measurement/digitizing scheme. Next, amplifier risetime effects on timing measurements are discussed, following by a brief tutorial on bandwidth and filtering effects on jitter measurements.

In discussing measurement accuracy and resolution, the authors list and describe the major components of error and how to successfully minimize their effect. In this section, amplitude error, aperture uncertainty, trigger jitter, interpolation error, and parameters for jitter measurement are defined. In discussing the parameters for jitter measurement, the note provides definitions for each parameter, including width, period, duty cycle, Delta t@level, Delta delay, Delta c2d+, Delta c2d–, frequency, and delay.

The application note concludes with the summing up of the principal actions for achieving accurate time measurement, such as maximizing the sample rate, using bandwidth-limiting filters, and using parameter-distribution histogram analysis. This application note is available as a free download from the company's website (www.lecroy.com).

LeCroy Corp., 700 Chestnut Ridge Rd., Chestnut Ridge, NY 10977-6499; (800) 553-2769, (845) 425-2000, FAX: (845) 425-8967, e-mail: corp@lecroy.com, Internet: www.lecroy.com.

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If the [signal] edges have equal risetime...limitation of the frequency can increase risetime without causing instability in the relative time of the signal edges.

Understand low-noise techniques for microwave oscillators

THE PRINCIPLE OF microwave oscillator design is based on the idea of negative-resistance generation as an attempt to compensate for resonator losses. One- and two-port oscillators provide several possible circuit combinations. In his 33-page application note entitled "Oscillator Basics and Low-Noise Techniques for Microwave Oscillators and VCOs," frequent *Microwaves & RF* contributor and Synergy Microwave Corp. (Paterson, NJ) Chairman Ulrich L. Rohde evaluates the conditions of oscillation for the Colpitts and Clapp-Gouriet oscillators. He then evaluates a 19-GHz silicon-germanium (SiGe)-based oscillator by assuming values that are backed up by available scattering (S)-parameters and DC I-V curves that were assigned to the nonlinear BFP520 model.

Rohde explains basic oscillator conditions, beginning with a short explanation of what an oscillator is and does, defining it as "an electronic circuit that overcomes the losses of a resonator by applying energy at the resonator frequency into the resonator." Once this is explained, he moves into his examination of the Colpitts oscillator, which is suitable for very-high-frequency (VHF) applications. Rohde then provides an introduction to the concept of phase noise, examines the nonlinear effects responsible for noise in oscillators, and offers insights into the ceramic-resonator-based oscillator (CRO). The paper resumes with an understanding of a lumped resonator oscillator (LRO) and investigates low-phase-noise sources.

Before discussing the microwave Clapp-Gouriet oscillator, Rohde provides a brief tutorial on distributed elements. By the conclusion of the application note, the reader has been treated to coverage of bipolar transistors and gallium-arsenide (GaAs) field-effect transistors (FETs). This application note is available as a free download from the company's website.

Synergy Microwave Corp., 201 McLean Blvd., Paterson, NJ 07504; (973) 881-8800, FAX: (973) 881-8361, e-mail: sales@synergymwave.com, Internet: www.synergymwave.com.

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cover story

Clock Translators Upgrade Data Links

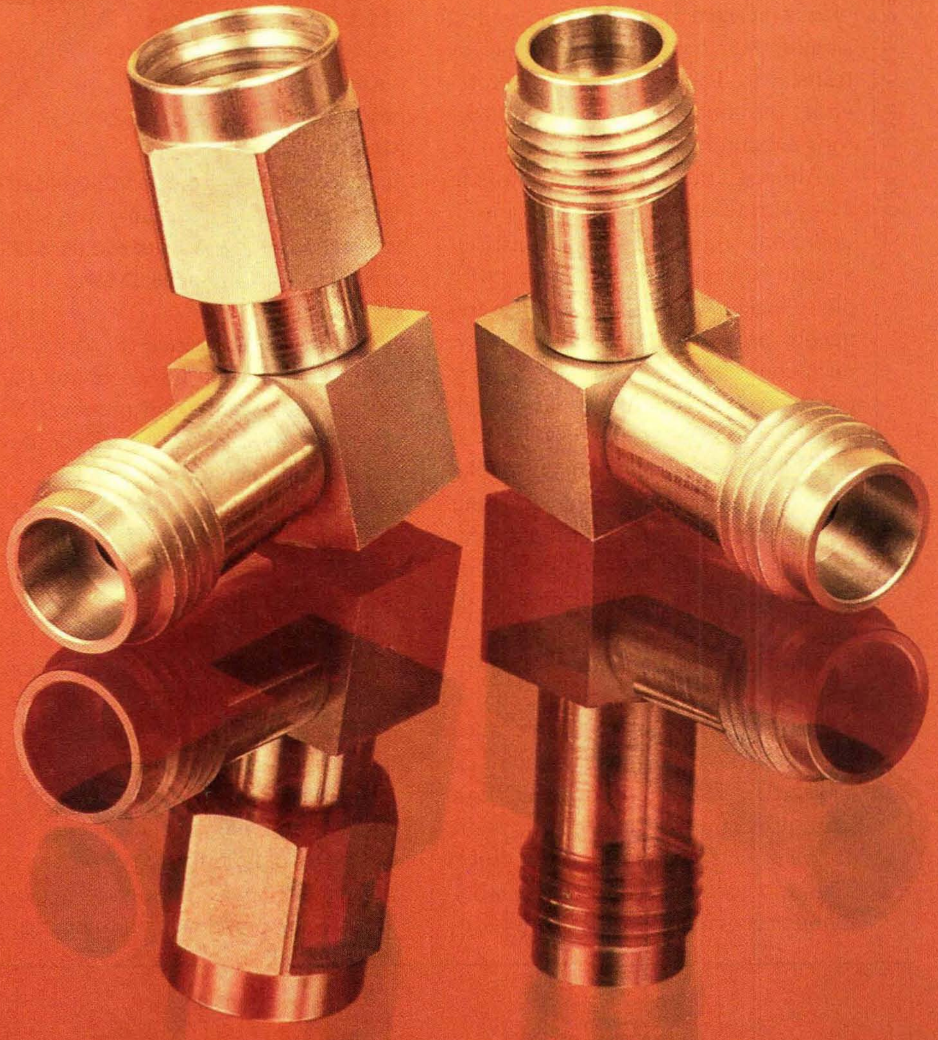
By leveraging low-noise VCSO technology, these stable clock translators shave jitter to a minimum while achieving excellent immunity to microphonics for output frequencies through 2488.32 MHz.

SHANKAR JOSHI
Chief Engineer
Synergy Microwave Corp., 201 McLean Blvd., Paterson, NJ 07504; (973) 881-8800, FAX: (973) 881-8361, e-mail: sales@synergymwave.com, Internet: www.synergymwave.com.

Clock translation is a vital function for upgrading the performance of digital communications systems, such as high-speed optical-carrier (OC) networks and Synchronous Optical Network (SONET) systems. The best clock-translation components, such as the new CTM-8-OCXX and CTS-8-OCXX series of devices from Synergy Microwave Corp. (Paterson, NJ), deliver precise output clock frequencies with very low phase noise (jitter). A key to the performance of these clock translators, which are available with standard output frequencies of 600.00, 622.08, 1244.16, and 2488.32 MHz, is the use of surface-acoustic-wave (SAW) technology in a line of low-noise voltage-controlled SAW oscillators (VCSOs). The clock translators provide a single sine-wave output that can be converted to positive-emitter-coupled-logic (PECL) and negative-emitter-coupled-logic (NECL) output formats.

For example, the model CTS-C1-12 clock translator is designed to generate an output frequency of 622.08 MHz when supplied with an input frequency of 51.84 MHz. For input signal levels of +10 dBm, the device delivers output signal levels to +6 dBm. It features input frequency-tracking capability of ±20 kHz and only suffers 3-ps typical jitter. The clock translator requires bias of 150 mA at +5 VDC and 10 mA at +8 VDC.

Another unit, the model CTS-C3-48, is designed to generate OC-48 output signals at 2488.32 MHz when supplied with a 155.52-MHz input signal. Input signals at a level of +10 dBm result in output signal level of +3 dBm. The clock translator features input frequency-tracking capability of ±20 kHz and typical jitter generation of 6 ps. The bias requirements are 150 mA at +5 VDC and 10 mA at +8 VDC. Both clock translators are designed for operating temperatures of –20 to +70°C.



The company offers clock translators **(Fig. 1)** designed to work with single (CTS-8-OCXX series) and multiple (CTM-8-OCXX) input signals. Multiple-input-frequency units can generate a precise output clock frequency for any of eight preselected input frequencies, using a multiplexer that is external to the clock translator. The multiplexer operates under the control of the clock-translator's integral microcontroller.

Although these clock translators fit a tiny surface-mount footprint of well-under one square inch, they consist of a fairly sophisticated subsystem, with phase-locked loop (PLL), reference-frequency preselector, decision circuitry, microcontroller, and VCSO **(Fig. 2)**. The preselector in the reference-frequency path ensures that reference input signals to the PLL are always less than 100 MHz. The preselector can pass reference input signals of less than 100 MHz unaltered, or divide them by a factor of two or three when the reference input signals are between 100 and 250 MHz.

The on-board microcontroller generates the clock, data, and latch-enable signals required to run the PLL integrated circuit (IC) contained within the clock translator. The microcontroller can store program information for eight different settings, supporting clock translation of up to eight different input signals. Provided with an input frequency, the microcontroller generates the appropriate data and command information to program the PLL's R- and N-counters to lock to the desired OC output signal. For example, to produce an OC-48 signal at 2488.32 MHz with a DSI reference input signal of 1.544 MHz, a common step size of 8 kHz (0.008 MHz) is used. In this case, the reference (R) division number is 1.544/0.008 = 193 and the main frequency-division number (N) is 2488.32/0.008 = 311040. Degradation of reference phase noise due to multiplication, according to the relationship 20logN, in this case is 20log(311040) = 109.856 dB.

1. The VCSO-based clock translators are available with standard output frequencies of 600.00, 622.08, 1244.16, and 2488.32 MHz.

Assuming a phase-detector noise floor of –160 dBm at the 8-kHz comparison frequency, the phase noise within the loop is thus –160 +109.856= –50.144 dBc/Hz. Depending upon the integrated-phase-noise requirements (normally from a few kHz to tens of MHz) of a specific application, the PLL's loop bandwidth can be adjusted for optimum loop performance. The integrated phase noise is dictated by the phase-noise envelope of the output signal over the required close-in bandwidth (offset from the carrier) and the spurious signals within the required bandwidth. If the spurious content can be controlled and minimized, the integrated phase noise can also be minimized.

For lower reference frequencies starting at approximately 8 kHz, the PLL filter loop bandwidth must be narrow (typically 1/20th of the step size) to reject the signals related to the step size. For the 8-kHz-step care, for example, the loop bandwidth must be approximately 400 Hz. In a typical PLL, this narrow loop bandwidth can lead to susceptibility to microphonics. But in these new clock-translation devices, the on-board VCSO features a rugged structure with good immunity to microphonics.

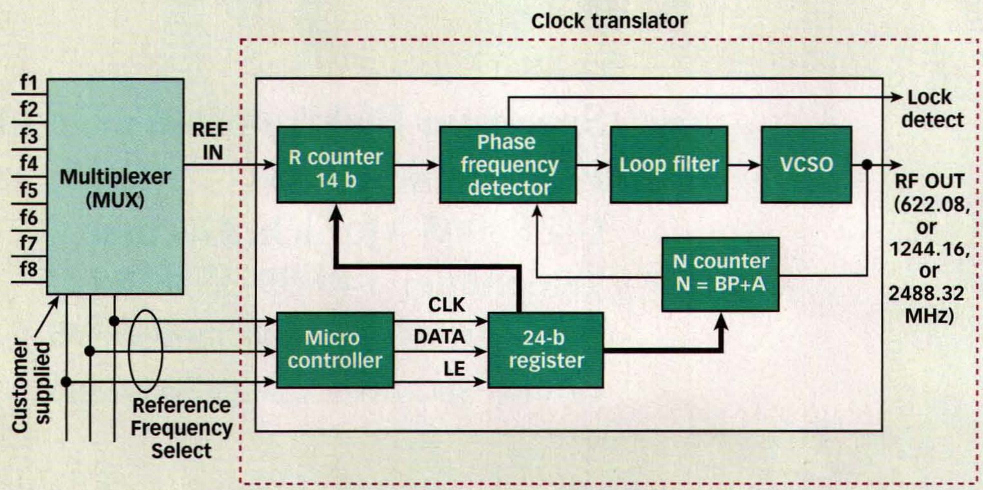
To achieve high performance in the PLL, the choice filter is critical. Closed-loop gain is a function of ($K_v \times K_d \times$ filter gain)/N,

where:

K_v = the VCSO constant and

K_d = the phase detector constant.

The filter gain must be kept high to



Input frequency	Select A B C	Example
f1 =	0 0 0	8 kHz
f2 =	0 0 1	64 kHz
f3 =	0 1 0	1.544 MHz
f4 =	0 1 1	2.048 MHz
f5 =	1 0 0	19.44 MHz
f6 =	1 0 1	44.736 MHz
f7 =	1 1 0	51.84 MHz
f8 =	1 1 1	155.52 MHz

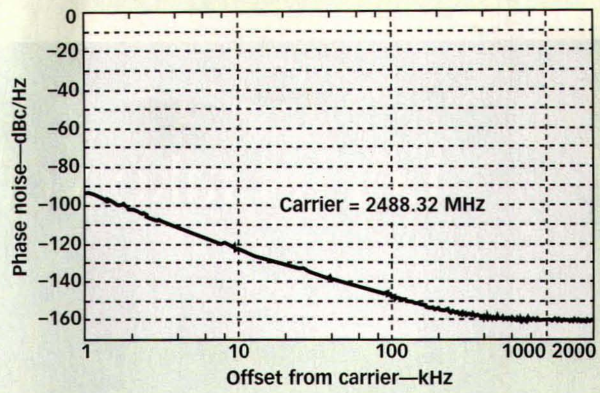
2. This block diagram shows the essential components of the clock translator, along with an external multiplexer for choosing among different input frequencies.

maintain loop stability under changing load and temperature conditions. The gain of a passive filter using a second-order loop with additional sections, for example, may be insufficient and active filters using operational amplifiers (opamps) may be required. The noise level of the opamp plays an important role in achieving high performance.

The new clock-translation components owe much of their stability and low noise to the performance of the integral VCSO. Much work has been performed over the last few years to improve this technology at higher frequencies (see *Microwaves & RF,* April 2001, p. 115). Although SAW resonators are difficult and expensive to manufacture at higher frequencies, the engineers at Synergy Microwave Corp. have developed a line of voltage-controlled crystal oscillators(VCXOs) through the use of frequency multiplication with output-frequency capabilities to 2488.32 MHz and beyond. Each VCSO consists of a SAW resonator along with a phase-shifting network in the feedback path of an amplifier. The overall insertion phase of the signal path is 0 deg. or multiples of 360 deg., with gain of greater than 1. Modern monolithic amplifiers are suitable as the active elements in these VCSOs. The VCSO's coupling and phase-shifting networks are designed for minimum loss and best impedance match with the SAW resonator.

The phase noise of the VCSO depends on the quality factor (Q) of the SAW resonator, the noise figure of the VCXO's active device (amplifier), the phase noise of the phase-shifter network, and the input and output coupling networks to the SAW resonator. The coupling networks should not degrade the resonator Q.

An example of an individual VCXO is the model VCSO-OC48, with a center frequency of 2488.32 MHz. It offers a tuning range of 250 PPM for tuning voltages of +1 to +4 VDC. It delivers typical output power of +3 dBm when operating with 60 mA at +5 VDC, and exhibits typical phase noise of –70 dBc/Hz offset 100 Hz from the carrier, –122 dBc/Hz offset 10 kHz from



3. This clock translator's phase noise was measured for various offsets from 2488.32 MHz.

the carrier, –142 dBc/Hz offset 100 kHz from the carrier, and –155 dBc/Hz offset 300 kHz from the carrier.

In each clock translator, the strength of the spurious signal corresponding to the step size and its harmonics depends on the quality of the phase detector. Spurious levels of –70 dBc or lower are not uncommon in modern designs. (When connecting a CTM-8-OCXX or CTS-OCXX module to surrounding circuitry, linear power supplies with enough supply decoupling should be used to achieve optimum performance.) The quality of the clock translator's output signal is a combination of the PLL and the VCSO performance. Outside the loop bandwidth, the VCSO dominates, whereas the performance of the PLL is controlled by the phase detector, frequency divider, and the external loop filter. **Figure 3** shows the phase noise of a 2488.32-MHz signal. The phase noise is –95 dBc/Hz offset 1 kHz from the carrier, –123 dBc/Hz offset 10 kHz from the carrier, –138 dBc/Hz offset 100 kHz from the carrier, and –160 dBc/Hz offset 1 MHz from the carrier.

In each clock translator, the strength of the spurious signal corresponding to the step size and its harmonics depends on the quality of the phase detector.

The jitter performance of the final clock translation output signal is determined by the quality of the VCSO and the PLL. Assuming worst-case spurious levels of –70 dBc, the root-mean-square (RMS) phase jitter is only 0.36 ps. The clock translators are suitable for SONET, local-multipoint-distribution-system (LMDS), and higher-order-modulation digital-radio applications. The tiny packaged devices measure only 0.625 × 0.512 × 0.031 in. (1.59 × 1.30 × 0.08 cm) and are well-suited for use with automated assembly equipment. The clock translators are available with standard output frequencies of 600.00, 622.08, 1244.16, and 2488.32 MHz, although other frequencies are available upon special request. The company also offers a coaxial test fixture for ease of evaluating any of the clock translators. **Synergy Microwave Corp., 201 McLean Blvd., Paterson, NJ 07504; (973) 881-8800, FAX: (973) 881-8361, e-mail: sales@synergymwave.com, Internet: www.synergymwave.com.**

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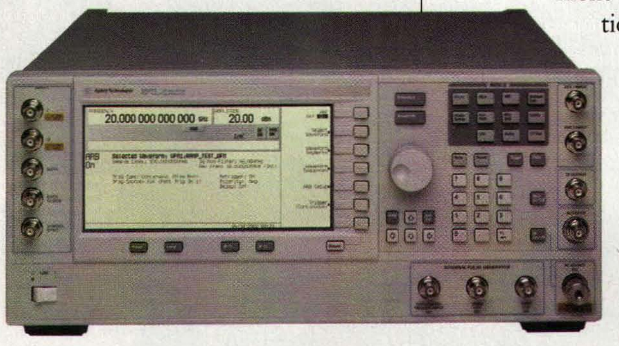
Signal Source Carries Vector Modulation To 20 GHz

The latest addition to a high-performance signal-generator line features complex modulation capabilities at carrier frequencies through 20 GHz.

JOHN HANSEN
Product Marketing Specialist, Electronic Product and Solutions Group
Agilent Technologies, 1400 Fountaingrove Pkwy., Santa Rosa, CA 95403.

Vector-modulated signal generators have been used to test digital communications systems since the advent of cellular's second generation (2G), approximately a decade ago. These sources have been limited to carrier frequencies of 6 GHz and below, until the introduction of the latest member of the Performance Signal Generator (PSG) line from Agilent Technologies (Santa Clara, CA). The Agilent E8267C extends the carrier range to 20 GHz, while also offering enhanced modulation and waveform-generation capabilities.

The Agilent E8267C **(see figure)** is the first single instrument to offer digitally modulated carriers above 6 GHz. The signal generator is characterized by outstanding phase-noise performance, fine frequency resolution, a wide output-power range, wideband in-phase/quadrature (I/Q) modulation capability, and a versatile built-in arbitrary-waveform generator (AWG) for creation of complex modulated waveforms. The instrument is suitable for traditional testing of wireless systems, such as cellular-communications components and networks, as well as higher-frequency applications such as radar, satellite communications (satcom), and broadband wireless-access systems.



The E8267C PSG Vector Signal Generator is the first test-signal source to offer digital modulation capabilities to 20 GHz.

The E8267C is based on a yttrium-iron-garnet (YIG)-tuned oscillator phase locked to a low-noise reference oscillator. The signal generator covers a total frequency range of 250 kHz to 20 GHz, with 0.001-Hz tuning resolution. It generates output levels up to +15 dBm, and can attenuate output levels down to –135 dBm. The signal generator features low phase noise of –110 dBc/Hz offset 10-kHz from a 10-GHz carrier and, with an option (UNB), can achieve even lower close-in phase noise of –104 dBc/Hz offset 500 Hz from a 10-GHz carrier.

To generate wideband and complex modulation formats, the E8267C allows operators to connect external analog I/Q signals to 160 MHz through BNC ports on the front panel. As an option (option 015), the signal generator can be equipped with wideband external I/Q ports that support uncorrected RF bandwidths as wide as 1 GHz. Complex modulation can also be generated internally, using an optional internal baseband generator based on a sophisticated dual-mode AWG. The internal baseband generator, which features a dedicated microprocessor, operates at a sample rate of 100 MHz. Since dual AWGs each achieve a better than 40-MHz Nyquist bandwidth at this sample rate, the resulting

I/Q modulation bandwidth of 80 MHz is more than enough to handle most modern modulation formats. The baseband generator incorporates 16-b digital-to-analog converters (DACs) running at 400 MHz for high sample playback resolution (0.001 Hz in frequency and 0.01 dB in amplitude).

The E8267C's generous 32-Msamples memory allows users to play back long, unique, and nonrepetitive waveforms. The large memory can reduce test time by facilitating measurements under multiple-signal environments with a single test waveform. Using the internal AWG, complex waveforms can be created by linking or sequencing a series of smaller, simpler waveforms. Up to 4096 waveform segments can be defined and sequenced to simplify the creation of custom waveforms. As an option (option 005), the E8267C is available with a 6-Gb hard disk for nonvolatile storage of waveform files.

The powerful AWG supports symbol rates from 1000 symbols/s to 50 Msymbols/s. The baseband generator operates at clock rates from 1 Hz to 100 MHz and achieves frequency resolution of 0.001 Hz. It can be used to generate a wide range of digital modulation formats, including binary phase-shift keying (BPSK), quadrature PSK (QPSK), offset QPSK (OQPSK), p/4-DQPSK, 8PSK, and 16PSK, minimum shift keying (MSK) for user-defined phase offsets from 0 to 100 deg., 4QAM to 256QAM, and frequency-shift keying (FSK) up to 16 deviation levels . The baseband generator can also be used to create custom I/Q maps consisting of as many as 256 discrete I and Q values. As far as digital filtering is concerned, the E8267C features a custom finite-impulse-response (FIR) filter with 16-b resolution that can support filter definitions to 64 symbols long. It is automatically resampled to a maximum length of 1024 coefficients, and is included with a variety of predefined digital filters, such as Nyquist, root Nyquist, and Gaussian filters.

Generating complex modulation is simple with the E8267C. Users can download their own arbitrary-waveform files or use the custom modulation capability to generate the desired signal. Modulation waveforms can also be defined in commercial software, such as MATLAB or Agilent's own Advanced Design System (ADS) suite of computer-aided engineering (CAE) tools, and downloaded to the signal generator. The company has also announced option 420 for the signal generator, which is the Pulse Builder Signal Studio software package. This allows radar-system developers to create optimized pulsed signals using the I/Q signals from the baseband generator.

Once a waveform file has been developed, operators can use a free program from Agilent known as the PSG/ESG Download Assistant to download the file into the E8267C. This application software, which works entirely within the MATLAB environment, allows operators to download MATLAB I/Q data into the volatile memory of the signal generator and perform playback with a single command. In addition to complex modulation, the E8267C PSG Vector Signal Generator can create two-tone and multi-tone signals for nonlinear device and component characterization. By pressing only a few softkeys on the signal generator, multi-tone waveforms can be created with precisely defined relative tone spacing, relative tone power, and phase relationships. Multiple signal sources and power-combining circuitry are no longer required, since a single E8267C can be used to create multi-tone signals at its RF output port.

The E8267C PSG Vector Signal Generator includes the ramp-sweep capabilities needed for compatibility with the company's 8757 series of scalar network analyzers. The signal generator is also available with an option (option 1E6) that supports the generation of narrow pulses (10 ns) at frequencies below 3.2 GHz. P&A: $66,700.00 (E8267C); 90 days. **Agilent Technologies, Test and Measurement Organization, 5301 Stevens Creek Blvd., MS 54LAK, Santa Clara, CA 95052; Internet: www. agilent.com/find/psg.**

Enter No. 52 at www.mwrf.com

Quad-Band PA Module Powers GSM And GPRS

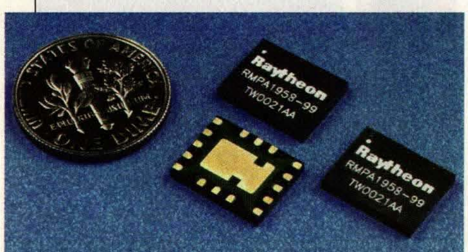
This versatile PA module offers analog power control and built-in input and output impedance matching across four different wireless-communications bands.

JACK BROWNE
Publisher/Editor

Saving space is definitely one of the design goals of the RMPA1958-99 quad-band power-amplifier (PA) module from Raytheon RF Components (Lowell, MA). Covering four different wireless bands from 824 to 1910 MHz, this robust PA module includes a band-select switch and analog power control to simplify the design of multimode, multiband cellular handsets for Global System for Mobile Communi-cations (GSM) and General Packet Radio Service (GPRS) applications.

The RMPA1958-99 supports the transmit functions of handsets intended not only GSM/GPRS systems, but also in Advanced Mobile Phone Service (AMPS), digital-communications-services (DCS), and personal-communications-services (PCS) systems.

The RMPA1958-99's total frequency coverage includes the US AMPS analog-cellular transmit band from 824 to 849 MHz, the GSM transmit band from 880 to 915 MHz, the DCS1800/DCS1900 transmit band from 1710 to 1785 MHz, and the PCS band from 1850 to 1910 MHz. Based on the company's proprietary indium-gallium-phosphide (InGaP) heterojunction-bipolar-transistor (HBT) technology, the compact module **(see figure)** features a complementary-metal-oxide-semiconductor (CMOS) analog power-control circuit, 50-Ω input and output matching, and internal DC blocking. The module achieves +35-dBm output power (for typical input power of +8 dBm) in GSM applications, +32.5-dBm output power (for +8-dBm input power) in DCS designs, and +31.5-dBm output power (for +8-dBm input power) in PCS applications.



The RMPA1958-99 PA module covers four different wireless bands with integral input and output impedance matching and analog power-control function.

The RMPA1958-99 quad-band GSM/GPRS PA module, which is designed to operate at a typical supply voltage of +3.5 VDC, features power-added efficiency (PAE) of 55 percent in US cellular GSM systems and 50 percent for DCS and PCS applications. The PA suffers very low leakage current of only 10 μA.

The RMPA1958-99 PA module selects bands through 0-to-+0.5-VDC control voltages for AMPS/GSM and +2.0 to +2.8 VDC for DCS/PCS. The power-control voltage ranges from +0.1 to +1.9 VDC with typically only 1-mA power-control current. The GSM/GPRS PA module is supplied in a leadless chip carrier that measures 11.6 × 9.1 × 1.6 mm to help efficiently use available handset printed-circuit-board (PCB) space. **Raytheon RF Components, 362 Lowell St., Andover, MA 01810; (978) 684-8900, FAX: (978) 684-5452, Internet: www.raytheonrf.com**

Enter No. 56 at www.mwrf.com

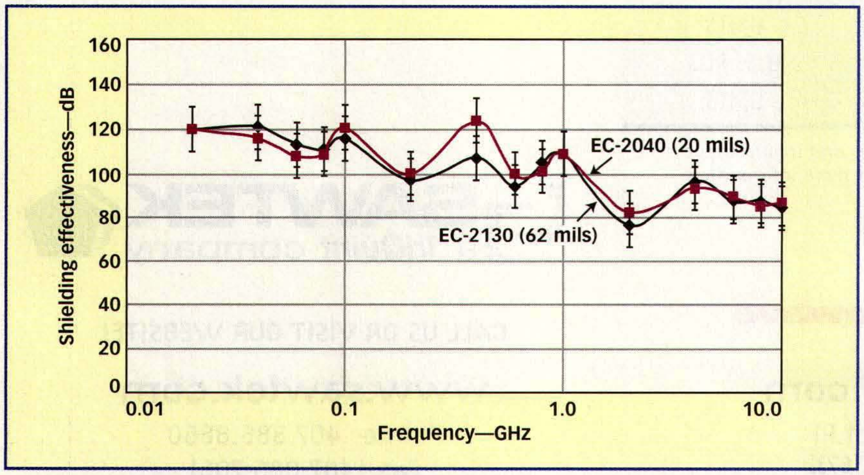
Conductive Silicones Supply Flexible Shielding

The BISCO EC-2000 series of conductive silicones are available in wide continuous rolls, supporting automated assembly for high-volume production runs.

JACK BROWNE
Publisher/Editor

Electromagnetic interference (EMI) has been a problem since the age of electronics began. Shielding against EMI is often the more desirable design approach because it is non-invasive to the circuitry. In addition, shielding has no effect on system speed while containing emissions and provides good immunity from outside sources. The growing need for EMI-shielding-capable gasketing is dictated by the trend toward increasingly higher operating frequencies (and shorter wavelengths) that result in decreased tolerances for even the smallest of gaps, seams, and slots in electronic equipment.

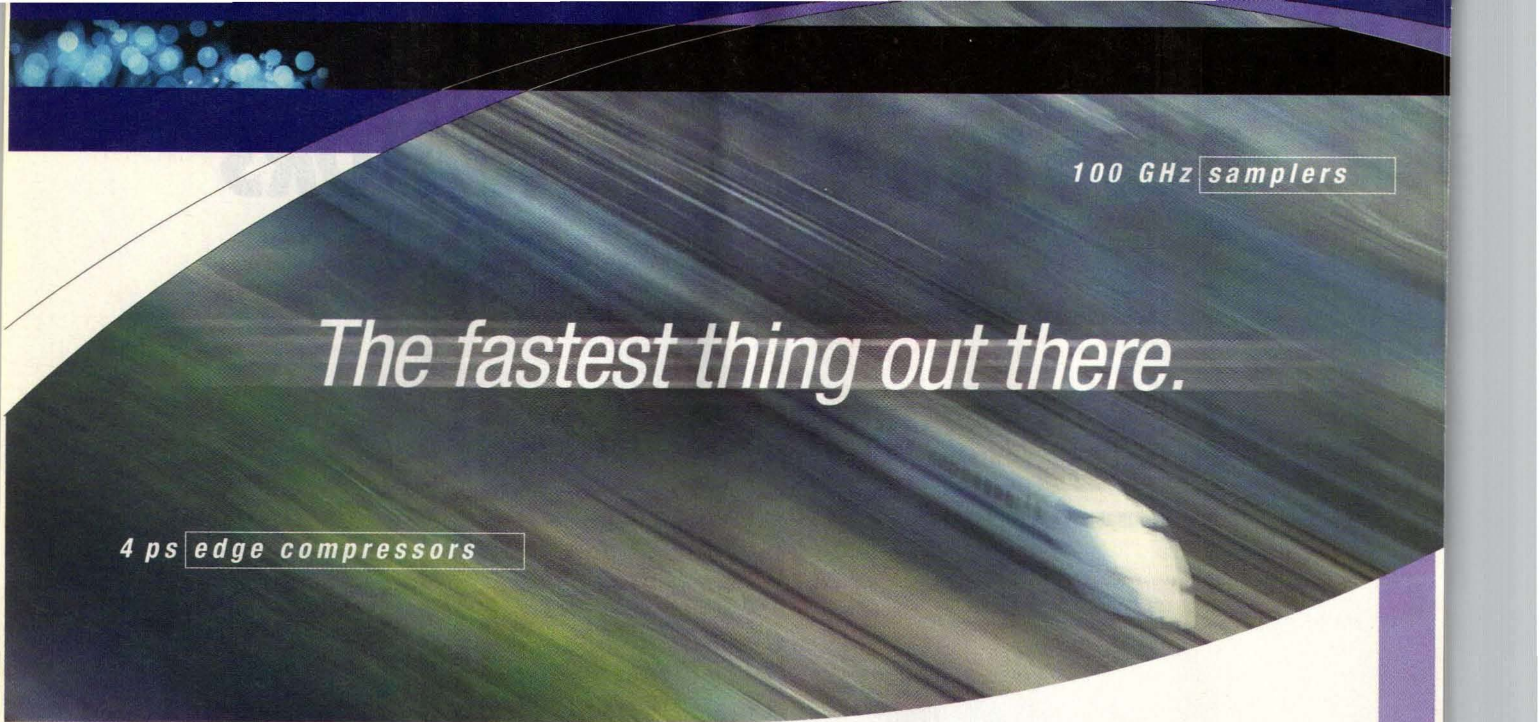
Within the wide selection of EMI gasketing technologies available, Rogers Corp. (Rogers, CT) is introducing its BISCO® EC-2000 Series Silicones, which offer high design flexibility by being much softer and more compressible than similar products on the market. This is significant with the current use of plastics and other lightweight enclosure materials that can withstand very little compressive force. With shielding effectiveness (SE) of better than 100 dB at 1 GHz, these thin, very soft, materials are supplied on wide continuous rolls to support shielding for large-volume devices, such as cellular telephones.



The shielding effectiveness of the EC-2040 and EC-2130 conductive silicones was evaluated according to MIL-G-83528 guidelines for frequencies from 10 MHz to 10 GHz.

The BISCO EC-2000 series materials are silicone resins filled with nickel (Ni)-coated graphite powder. These materials carry a Underwriters' Laboratories (UL) 94V-1 (vertical) flame resistance rating at most thicknesses, with an 94-HB horizontal flame-resistance rating on all thicknesses. The company currently offers EC-2040 material in thickness from 0.50 to 1.59 mm at 40 Shore A Durometer and EC-2130 material in thickness from 1.59 to 3.18 mm at 30 Shore A Durometer. The materials achieve SE performance far exceeding 100 dB at lower frequencies (10 MHz to 1 GHz), while still maintaining shielding of 85 dB at frequencies up to 10 GHz **(see figure).**

Increasing use of RF energy for communications—in cellular systems, wire-



The creator of the industry's fastest pulse generators introduces two new technologies to its diverse family of products.

Our two new product offerings are examples of Picosecond's 20 year commitment to innovation, reliability and performance. These revolutionary advancements are setting new standards for high-speed signal generation and acquisition.

4 ps EDGE COMPRESSORS

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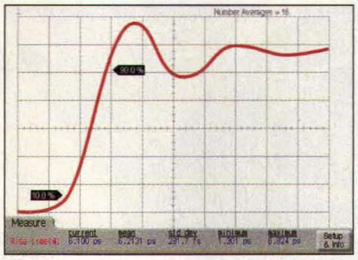
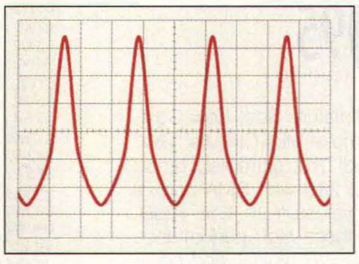


EDGE COMPRESSOR OUTPUT
Input: 7 Vp-p, 34 ps risetime. Estimated system risetime: 5.7 ps. Measured risetime: 6.1 ps. Estimated signal risetime: < 4 ps.



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Input: 10 GHZ, 6V p-p sine wave. Output: 3V p-p, 22 ps. (FWHM).

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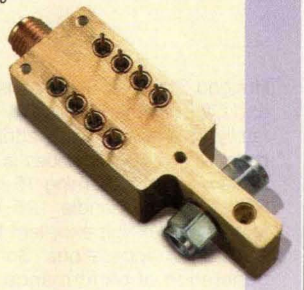
- *Ultra-fast metrology*



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less local-area networks (WLANs), and Bluetooth devices—heightens the need for controlling radiated emissions at RF and microwave frequencies. At the same time, these communications devices must be designed to be relatively immune to external sources of RF and microwave radiation. Bluetooth and WLAN systems operate in the unlicensed frequency band from 2.400 to 2.483 GHz. But commercial and consumer microwave ovens also operate within that band, specifically at 2.450 GHz. And with the increasing clock rates of personal computers (PCs) already encroaching on the frequency bands reserved for cellular and personal-communications-services (PCS) equipment, it is easy to see the enhanced role of shielding materials, such as the EC-2000 conductive silicones, in the design of RF- and microwave-based devices. In addition to assisting the desired operation of electronic devices, shielding also plays an important role in minimizing exposure to excessive levels of EM radiation. (For more on this important topic, please refer to the article by A. Kumar on p. 72.)

EM radiation occurs close to (the near-field radiation region) and farther away from (the far-field radiation region) from a source, such as a cellular telephone's local oscillator (LO). As the operating frequency increases, the transition point between the radiated energy's near and far fields moves closer to the source. In developing a shielding solution, the ideal approach is to contain the EMI as close to the source as possible, minimizing far-field effects.

When a combination of high current and low voltage are used in a circuit, a high magnetic field and low electric field results, as found around motors and transformers. On the other hand, the combination of high voltage and low current yields a high electric field and low magnetic field, as typified by the clock oscillator in a computer. The low-voltage/high-current combination produces a low wave impedance, while the high-voltage/low-current combination yields a high wave impedance. By using a shield with significant impedance mismatch to the radiated wave impedance, the impedance wave can be reflected away from the direction of propagation, and effectively attenuated. To do this, a high-current source requires a magnetic permeable shield, while a high-voltage source requires an electrically conductive shield.

The effectiveness of an EMI shield is determined by how EM waves are reflected by the surface of the shield, how the incident EM waves are absorbed by the bulk material in the shield, and how the EM waves are re-reflected from the back surface. The re-reflected energy can add or subtract from the SE, depending on the phase relationship of the reflected waves with the incident waves. A material's SE is related to the thickness of the shielding material, the relative conductivity of the material, the relative permeability of the material, and the frequency of the incident EM radiation. For many applications, SE levels approaching 60 dB are considered good, levels well-exceeded by the minimum performance levels of the thinnest of EC-2000 materials.

To briefly compare the two EC-2000 materials, the EC-2040 material has a specific gravity of 2.18, while the EC-2130 material has a specific gravity of 1.97. The EC-2040 material features minimum tensile strength of 90 psi, minimum elongation of 60 percent, volume resistivity of 0.2 Ωcm, and SE levels [for a 0.020-in. (0.051 cm)-thick material] of 120 dB at 100 MHz, 120 dB at 500 MHz, 110 dB at 1 GHz, and 85 dB at 10 GHz. The thicker EC-2130 material offers minimum tensile strength of 50 psi, minimum elongation of 50 percent, volume resistivity of 0.7 Ωcm, and roughly comparable SE levels of 100 dB at 100 MHz, 100 dB at 500 MHz, 110 dB at 1 GHz, and 85 dB at 10 GHz.

The BISCO EC-2000 series silicones feature unsurpassed compression-set resistance and SE performance that belies the thin, soft nature of these materials. **Rogers Corp., One Technology Dr., P.O. Box 188, Rogers, CT 06263-0188; (860) 774-9605, FAX: (860) 779-5509, Internet: www.rogerscorporation.com.** MRF

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ASIC Stabilizes Crystal Oscillators

A unique circuit applies fourth-order temperature compensation to surface-mount VCXOs for stability up to ±0.3 PPM.

JACK BROWNE
Publisher/Editor

Crystal oscillators are the frequency reference sources in any communications system. But achieving high stability usually requires some form of temperature compensation, to adjust for the tendency of quartz resonators to shift in frequency as a function of temperature. By applying analog temperature compensation by means of a proprietary application-specific integrated circuit (ASIC), however, the engineers at C-MAC MicroTechnology (Durham, NC) have developed a line of extremely stable miniature temperature-compensated, voltage-controlled crystal oscillators (TCVCXOs) with minimal current consumption.

The company's CFPT-9000 and CFPT-9050 series TCVCXO models are typically available with output frequencies from 8 to 50 MHz. Standard output signals are clipped sine waves, with peak-to-peak output levels of typically +1.2 VDC, although the oscillators can be used to generate standard sine waves, as well as square high-performance complementary metal-oxide semiconductor (HCMOS) and square analog-CMOS (ACMOS) waveforms. The crystal oscillators can be specified for linear voltage control of frequency from ±5 to ±50 PPM. The phase noise for both series is outstanding, at −95 dBc/Hz offset 10 Hz from a 13-MHz carrier and −145 dBc/Hz offset 100 kHz from the carrier.

The compact TCVCXOs combine proven crystal-oscillator circuitry with the company's "Pluto" ASIC for temperature compensation. The CFPT-9000 and CFPT-9050 series TCVCXOs are capable of better than ±0.3-PPM frequency stability from −20 to +70°C. The aging rates for both oscillator series are ±1 PPM in the first year and ±3 PPM for 10 years.

The oscillators can operate on power supplies as low as +2.5 VDC. Current consumption is as low as 1.5 mA. Their low jitter make them suitable as Statum 3 clocks in SONET and SDH systems.

The TCVCXOs can be ordered with different levels of frequency stability depending upon temperature, with the ±0.3-PPM performance available from 0 to +50°C and 0 to +70°C; stability of ±1 PPM is available from −40 to +85°C. The CFPT-9000 series is supplied in a 7 × 5 × 2-mm ceramic surface-mount package while the CFPT-9050 series is available in a 14.0 × 9.0 × 6.1-mm SOJ-20 package. **C-MAC MicroTechnology, 4222 Emperor Blvd., Suite 300, Durham, NC 27703; (919) 474-3500, FAX: (919) 941-0530, e-mail: cmac.microtechnology@solectron.com, Internet: www.cmac.com/mt/.**

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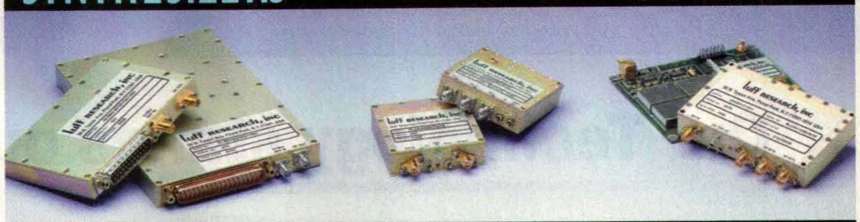

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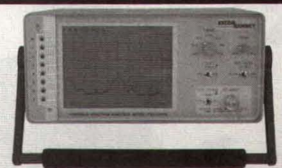

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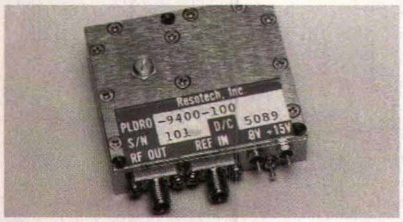

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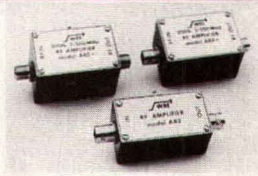

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MODEL 430803E IS a 26-GHz single-pole, six-throw (SP6T) switch with internal 50-Ω termination. The switch operates on voltages of +24 to +32 VDC (+28 VDC nominal) with typical current draw of 180 mA. Equipped with a latching actuator, the component switches

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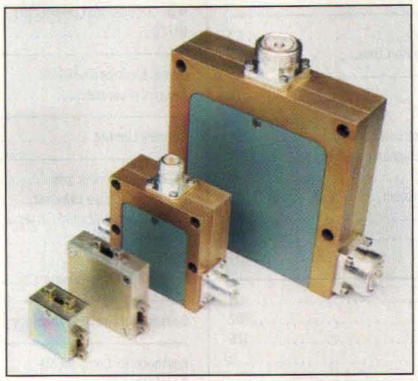

CW, and 2 kW peak in the 400-to-800-MHz television bands where 250-W stripline units are available. In the 0.8-to-3.5-GHz spectrum, 0.15-dB loss stripline units operate at 200-W CW, 2-kW peak power levels.

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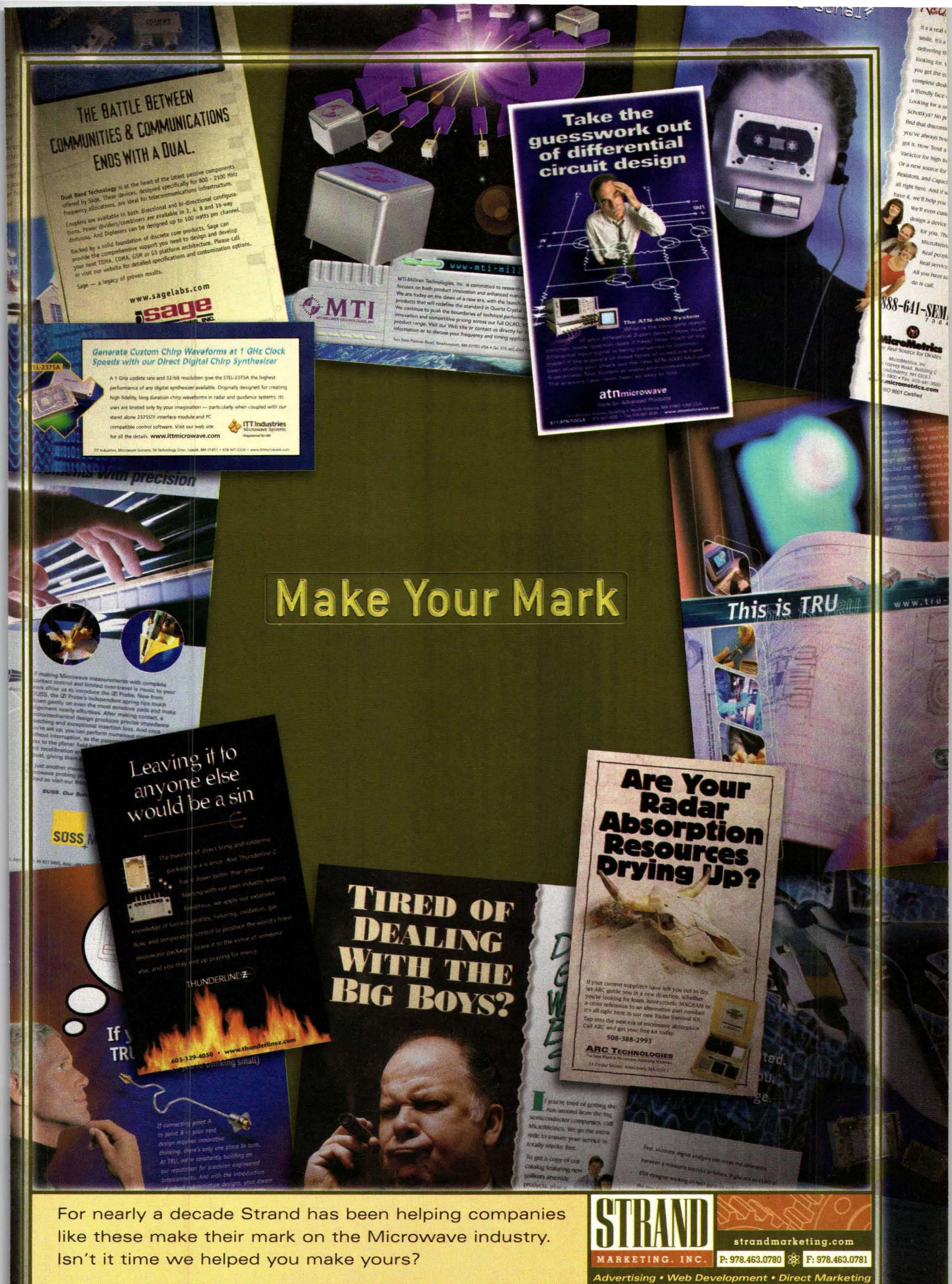

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looking back

NEARLY 30 YEARS ago, a miniature amplifier developed by Varian Associates (Palo Alto, CA) achieved 10-mW output power from 2 to 4 GHz at the then-unheard-of size of only $0.5 \times 1.0 \times 1.76$ in. (excluding connectors). The octave-band amplifier was designed for communications and ECM systems.

next month

Microwaves & RF October Editorial Preview

Issue Theme: Integrated Circuits

News

October highlights emerging technologies in ICs, detailing not only process refinements and circuit advances, but also the applications for these newer technologies. The Special Report will provide updates on the latest HBT and HEMT devices (with a sneak preview of papers to be presented at the upcoming International Electron Devices Meeting), as well as a rundown on improvements in GaN, SiC, and SiGe processes. The report will include how the latest devices are being used in such applications as cellular Txs, WLANs, and UWB systems.

Design Features

The October issue will include several technical contributions on IC designs, including a single-chip direct-conversion Rx IC developed by a leading semiconductor supplier. The device is designed to directly demodulate signals in the frequency range from 800 to 2700 MHz. Additional technical articles highlight a single-chip UHF transceiver designed for low-voltage, low-current operation, the use of a powerful design and verification software suite to simplify the design of dense ICs, and techniques for applying balanced error correction to improve the linearity of high-power amplifiers.

Product Technology

The October Product Technology will introduce a series of low-profile (two-rack-space) vector-modulated signal generators for emulating complex modulation formats in communications systems through 4 GHz. These novel signal sources offer touch-screen control and a built-in AWG to create all of the digital-modulation formats commonly used in wireless-communications systems. Additional Product Features will feature a line of high-performance signal-acquisition instruments, a VNA system that works with modulated test signals, and a family of SiGe-based devices for GPS applications.